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Television Applications of the PFL200

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The Mullard PFL200 double-pentode has many applications in receivers designed for negative or positive vision modulation, and in dual-standard receivers. This article describes video amplifier, a.g.c., sync pulse separator, and black-level stabilisation circuits.

INTRODUCTION

A simplified dual-standard video detector and a.g.c. system using the PFL200 has already been described (Ref. 1). Further applications of this new double-pentode are now given.

PFL200 OUTPUT SECTION

The PFL200 'L' section is designed to operate as a high-gain video output pentode capable of producing a large output voltage across a low-value anode load resistor. To achieve high gain, which also permits the application of negative feedback, a frame-grid control grid has been used, and this gives the valve a slope of approximately 21mA/V at 30mA. The high gain of this section will also make the valve particularly useful in hybrid television receivers in which the last i.f. amplifier is transistorised.

The design of the 'L' section has taken note of the increasing popularity of high-level contrast control circuits in which the contrast control potentiometer is placed in the anode circuit of the video output valve. Such circuits provide constant drive to the synchronising pulse separator and a.g.c. system regardless of the setting of the contrast control. The capacitance added to the anode when this technique is employed necessitates the use of a low-value anode load resistor. The PFL200 can provide adequate current to produce a composite peak-to-peak video output voltage of 100V across an anode load as low as 2kΩ.

Care has also been taken in the valve design to permit screen-grid dissipation limits that allow for short term overloading. This can arise when a tuner in a television receiver operating from a signal having negative modulation is switched so that neither video signal nor noise is present to provide bias for the grid of the video output valve.

PFL200 AMPLIFIER SECTION

The 'F' section of the PFL200 is a voltage-amplifying medium-slope pentode designed for auxiliary functions of the receiver. In receivers intended for the U.K. market, it is most likely to find application as a synchronising pulse separator, and the valve has been designed to provide adequate current at a low anode voltage for this application. In receivers intended for overseas markets, where synchronising pulse separation is usually effected using an interference-gated circuit employing a heptode valve, the 'F' section of the PFL200 can be employed with advantage as a high-gain a.g.c. amplifier, or as an intercarrier sound i.f. amplifier. To facilitate these applications, care has been taken to reduce to a minimum the feedback capacitance from the anode of the 'F' section to other electrodes in the double valve.

Applications in Receivers Designed for Negative Vision Modulation

VIDEO AMPLIFIER CIRCUITS

Circuits with Large Load Capacitance

Designed to accommodate large shunt capacitances in the anode circuit, such as those resulting from the use of high-level contrast controls, the PFL200 output section can provide a 50mA signal current swing, thus producing a peak-to-peak output of 100V with a 2kΩ anode load.

To avoid exceeding the anode dissipation limit of 5W, it is necessary to keep the mean anode supply potential $V_{b(max)}$ to a value such that

$$(0{\cdot}5\ V_{b(max)})^2/2000 = 5W;$$

that is to say $V_{b(max)}$ is 200V. For operation from h.t. lines in excess of 200V, a decoupled resistor should be included in the anode supply.

Some resistance in the cathode circuit is desirable to prevent the valve working too close to the positive grid region on signal peaks, when the onset of grid current might load the video detector circuit and distort the signal. In addition, the cathode resistor also assists in controlling the no-signal operating conditions, as well as providing means for h.f. compensation.

With a cathode resistor, the high mutual conductance of the valve enables a stage gain of some 30 times to be achieved. Gains in excess of this figure can be realised, but in a practical receiver non-linearity of the video detector under conditions of low contrast may preclude their use.

A resistor is included in the screen-grid circuit to prevent excessive dissipation under no-signal conditions. To retain the full d.c. component of the signal this resistor should not be decoupled, although an h.f. bypass capacitor is desirable to minimise Miller effect (dotted in Fig. 1).

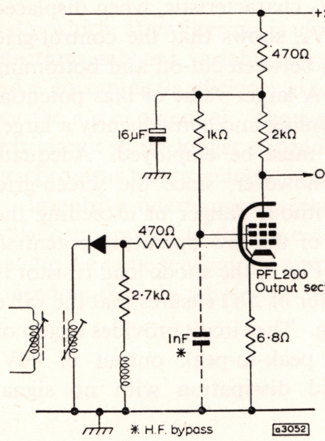


Fig. 1—Video amplifier with 2kΩ anode load

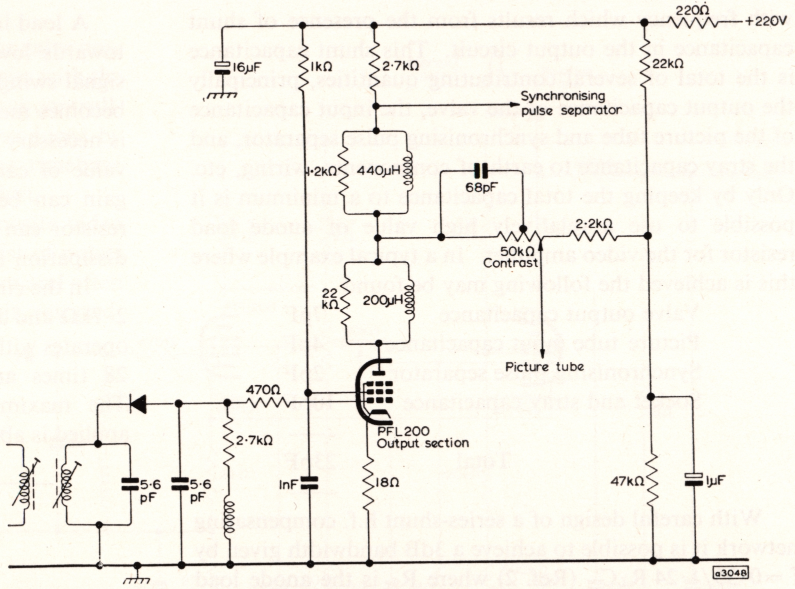


Fig. 2—Video amplifier with high-level contrast control

The resistor value must be chosen in conjunction with that of the cathode resistor to establish the desired stage gain.

In the circuit shown in Fig. 1 (R_L = 2kΩ, R_k = 6·8Ω, R_{g2} = 1kΩ undecoupled), a maximum peak-to-peak output signal of about 120V can be produced from an input signal of 3·6V, giving an overall gain of 33 times.

The use of a high-level contrast control is illustrated in the extract from a complete receiver circuit shown in Fig. 2. The video amplifier operates at maximum signal level at all times, an advantage for synchronising pulse separator action. The video signal to the picture tube is taken from a potentiometer forming the contrast control in the video amplifier anode circuit. In order not to change the picture black level with contrast control operation, the end of the potentiometer remote from the anode is connected to a d.c. potential equal to the black-level potential at the video amplifier anode. During black elements of the picture there is then no potential difference across the potentiometer, and the slider will be at black-level potential whatever the setting.

The series resistance of the contrast control circuit preceding the input capacitance of the picture tube results in a loss at the higher video frequencies unless compensation is provided. Such compensation can take the form of capacitive bypassing of one or more sections of the potentiometer track. Potentiometers are commercially available for this purpose with taps at either 33%, 50% or 66% rotation. For most purposes compensation at only one tapping point, 66%, is satisfactory, since this is the most common setting of the contrast control. Furthermore, compensation will be approximately correct for a range of settings on either side of this point.

A high value of resistance is desirable for the contrast control to minimise dissipation. The mounting of the control should be such as to minimise the capacitance to earth. The circuit shown in Fig. 2 has been incorporated in an experimental receiver with very satisfactory results, the control being mounted on an insulated panel. The value of the compensating capacitor must be chosen when the final layout has been established.

Circuits for Direct Operation from 220V H.T. Potential

Where the shunt capacitance permits the use of higher values of anode load, operation directly from a 220V h.t. line is possible. The minimum value of anode load to ensure operation within the permitted anode dissipation limit of 5W is given by $R_L = (0{\cdot}5 \times 220)^2/5 = 2424\Omega$.

A practical circuit of a video amplifier with an anode load resistor of 2·7kΩ is shown in Fig. 3. In this circuit, a gain of about 30 times may be achieved with a reduced effective slope for the valve, and for this reason a cathode resistor of 18Ω has been selected.

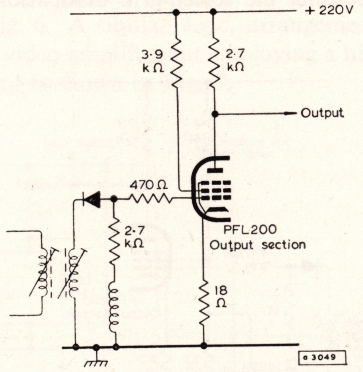


Fig. 3—Video amplifier with 2·7kΩ anode load

The higher value of anode load causes the valve to operate further below the knee in the no-signal condition, with a resulting increase in screen-grid current. Consequently, a larger value of screen-grid resistor must be employed to ensure that the maximum screen-grid dissipation is not exceeded. In the circuit in Fig. 3 the screen-grid resistor is 3·9kΩ. With this circuit (R_L = 2·7kΩ, R_k = 18Ω, R_{g2} = 3·9kΩ undecoupled), a maximum video output signal of about 120V is produced with an input signal of 3·6V.

Circuits with High Value of Anode Load Resistance

In a video amplifier circuit it is generally necessary to provide means of compensation for the falling response with frequency which results from the presence of shunt capacitance in the output circuit. This shunt capacitance is the total of several contributing quantities, principally the output capacitance of the valve, the input capacitance of the picture tube and synchronising pulse separator, and the stray capacitance to earth of components, wiring, etc. Only by keeping the total capacitance to a minimum is it possible to use a relatively high value of anode load resistor for the video amplifier. In a typical example where this is achieved the following may be found

Valve output capacitance	7pF
Picture tube input capacitance	4pF
Synchronising pulse separator	2pF
Socket and stray capacitance	10pF
Total	23pF

With careful design of a series-shunt h.f. compensating network it is possible to achieve a 3dB bandwidth given by $f = 0\cdot35/1\cdot24\, R_LC$ (Ref. 2) where R_L is the anode load resistance and C is the total shunt capacitance. Thus, for a bandwidth of 4Mc/s, with 23pF shunt capacitance:

$$R_{L(max)} = \frac{0\cdot35}{1\cdot24 \times 4 \times 10^6 \times 23 \times 10^{-12}} = 3060\Omega.$$

A video amplifier circuit has been investigated using a 3·3kΩ anode load (without h.f. compensation). The circuit diagram is shown in Fig. 4. As described in the preceding section, the cathode resistor value has been increased to establish the desired gain, and in this case it is 27Ω. The screen-grid circuit incorporates a potentiometer for contrast adjustment via the a.g.c. system. A minimum value of 4·7kΩ in the screen-grid supply is necessary to avoid exceeding the screen-grid dissipation limit.

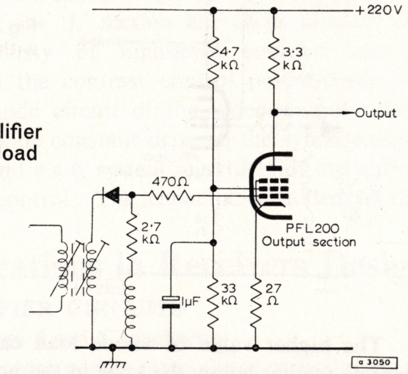


Fig. 4—Video amplifier with 3·3kΩ anode load

In the circuit of Fig. 4 ($R_L = 3\cdot3k\Omega$, $R_k = 27\Omega$, $R_{g2} = 4\cdot7k\Omega + 33k\Omega$, decoupled) at maximum contrast, a peak-to-peak video output signal of 110V is produced with an input of 3·6V.

Circuits for Operation from Lower H.T. Potentials

In some countries, notably the U.S.A. and Japan, the supply mains potential is 117V. In many receiver designs a half-wave rectifier is employed for the h.t. supply, the resulting h.t. potential being about 130V. The characteristics of the PFL200 enable a large video signal to be obtained when operating in this type of receiver.

A load line on the I_a/V_a characteristic, when displaced towards lower values of V_a, shows that the control-grid signal swing for operation between cut-off and bottoming becomes more restricted. A larger value of bias potential is necessary to avoid bottoming and consequently a larger value of cathode resistor must be employed. Adequate gain can be maintained, however, since the screen-grid resistor can be omitted without danger of exceeding the dissipation limit, because of the low operating potential.

In the circuit shown in Fig. 5, the anode load resistor is 2·7kΩ and a cathode resistor of 27Ω ensures that the valve operates without bottoming. The circuit provides a gain of 28 times and maximum peak-to-peak output of 80V. The maximum screen-grid dissipation with no signal applied is about 1·5W.

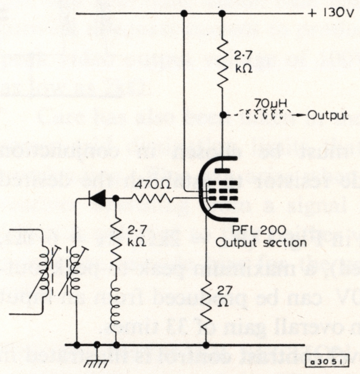


Fig. 5—Video amplifier with 130V anode supply

The transmitted video bandwidth of the 525-line system is 4Mc/s, and hence the bandwidth of a video amplifier employing a 2·7kΩ anode load is adequate with very simple h.f. compensation. With a load capacitance of 25pF, the uncompensated bandwidth is about 2·5Mc/s. The bandwidth can be increased to about 4Mc/s with a series compensating inductance in the anode circuit, as shown dotted in Fig. 5. The use of a higher value of anode load resistance does not result in an increase of gain or of maximum output since these are offset by the corresponding increase in the cathode resistor value necessary to avoid bottoming.

CIRCUITS USING VOLTAGE AMPLIFIER SECTION

Interference-protected A.G.C. Circuit

A number of circuits for deriving a.g.c. potentials in receivers for negative modulation make use of line flyback pulses (Ref. 3). In the simplest form, positive-going pulses are a.c.-coupled to the anode of a valve functioning as a controlled d.c. restorer, the video signal being d.c.-coupled to the grid or cathode. The valve responds to the d.c. level of synchronising pulse tips and governs the mean negative potential of the d.c.-restored line pulses at the anode. After filtering, this negative potential is used for a.g.c. This arrangement has a certain inherent immunity to interference because it is inactive between the line flyback pulses. It suffers from the disadvantage, however, of being dependent on line timebase synchronism. To overcome this disadvantage a voltage-dependent resistor (v.d.r.) is included in the anode circuit (Ref. 4). The non-linear characteristic of the v.d.r. and the asymmetric waveform of the flyback pulses provide a d.c. component across the v.d.r. which, after filtering, is used for a.g.c. The valve now functions only as a variable d.c. resistance in series with the v.d.r. governing its efficiency as a rectifier and thus controlling the a.g.c. potential. The a.g.c. potential produced by this circuit is no longer dependent on the line timebase synchronism but immunity to interference is lost.

Impulsive interference present in the composite video waveform applied to the a.g.c. valve will consist of pulses having the same sense as the synchronising pulses and often exceeding them in amplitude. The a.g.c. circuit, responding to the level of synchronising pulse tips, will also be actuated by the interference pulses, which create a disturbance of the control potential. Owing to the relatively long time-constant of the a.g.c. circuit the effects of interference persist longer than the duration of the interference itself. The circuit can be protected against the effects of interference by suitably gating the a.g.c. valve, as is done with synchronising pulse separators. For this purpose it is necessary to use the control grid, which involves applying the composite video signal to an alternative electrode (for example, the screen grid, Ref. 4).

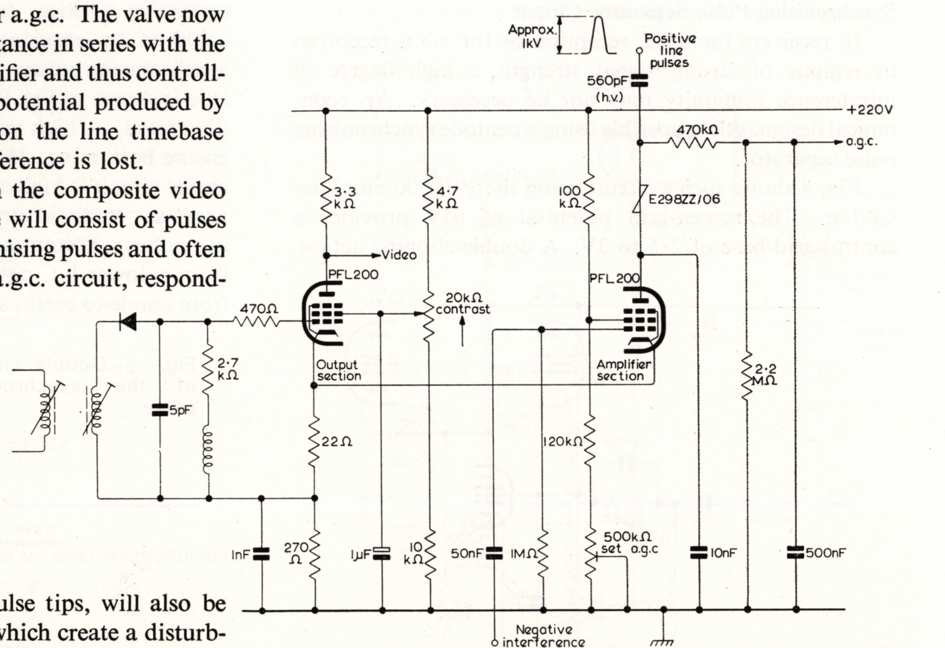


Fig. 6—Video amplifier and synchronising-tip a.g.c. circuit with low-level contrast control

An alternative circuit technique is to apply the video signal to the cathode from a low impedance. This can readily be done by connecting the cathode to the video amplifier cathode, and employing a common resistor. However, the d.c. gain of the video amplifier is no longer in the a.g.c. loop, and the circuit with a triode a.g.c. valve may not be sufficiently sensitive. With the use of a pentode as the a.g.c. valve, the loop gain is adequate and interference gating pulses, generally already available in the receiver for the protection of the synchronising pulse separator, can be applied to the control grid. These pulses are negative-going and hold the valve cut off for the duration of the interference, thus preventing any disturbance of the control potential. A complete circuit is shown in Fig. 6. A similar a.g.c. arrangement in conjuction with a video amplifier but employing a high-level contrast control is shown in Fig. 7.

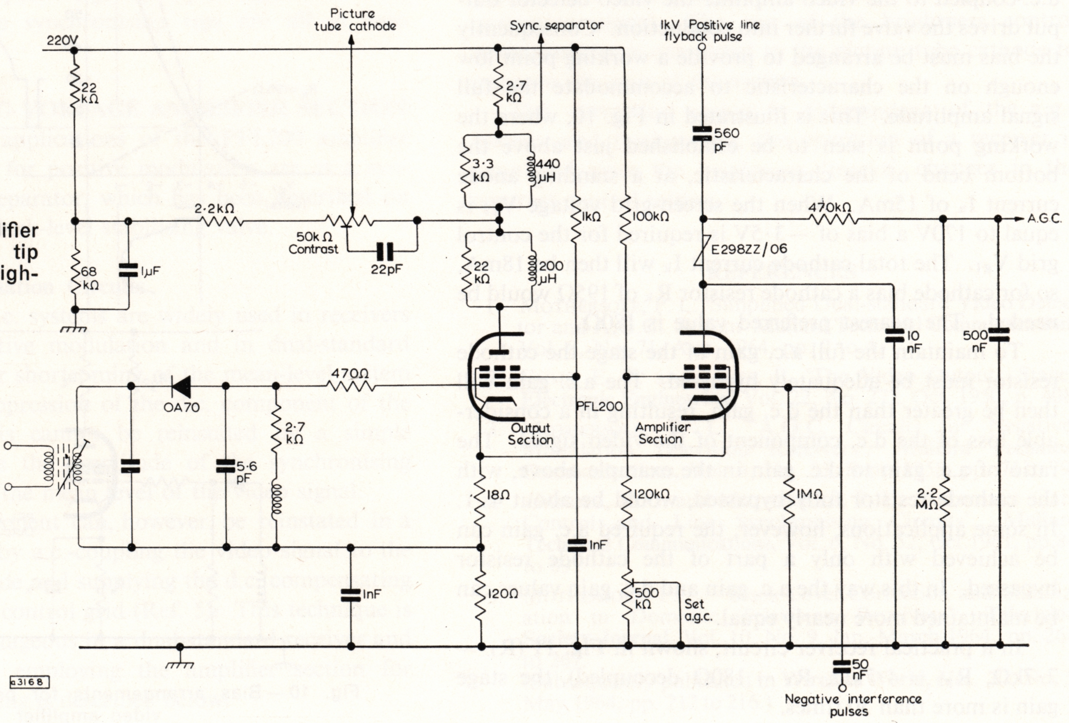


Fig. 7—Video amplifier and synchronising tip a.g.c. circuit with high-level contrast control

Synchronising Pulse Separator Circuit

In receivers for u.h.f. reception or for v.h.f. reception in regions of strong signal strength, a high degree of interference immunity may not be necessary. An economical design is then possible using a pentode synchronising pulse separator.

Fig. 8 shows such a circuit using the PFL200 amplifier section. The screen-grid potential of 65V provides a control-grid-base of 2·5 to 3V. A double-clipping action, extracting a "slice" from the synchronising pulse, can be achieved by ensuring that the valve bottoms fully on synchronising pulses. The grid-to-anode transfer characteristic is shown in Fig. 9 where the double-clipping action is illustrated. A high value of anode load resistor is used to ensure bottoming. However, if the load capacitance in the anode circuit is high a smaller value of anode load may be necessary to avoid stretching of the separated synchronising pulses; bottoming can then be assured by operation from a lower h.t. potential, achieved by a shunt resistor from anode to earth, shown dotted in Fig. 8.

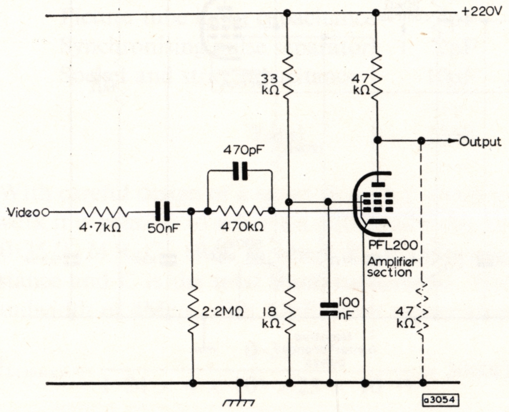
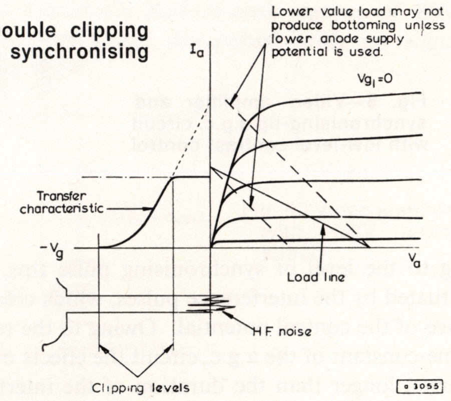


Fig. 8—Synchronising pulse separator

Fig. 9—Double clipping of the synchronising pulse



Applications in Receivers Designed for Positive Vision Modulation

VIDEO AMPLIFIER CIRCUITS

Video amplifier circuits for positive modulation differ from their negative modulation counterparts in the bias arrangements for the valve. In this application the video detector output is positive with respect to earth. When d.c.-coupled to the video amplifier the video detector output drives the valve further into conduction. Consequently the bias must be arranged to provide a working point low enough on the characteristic to accommodate the full signal amplitude. This is illustrated in Fig. 10, where the working point is seen to be established just above the bottom bend of the characteristic, at a standing anode current I_a of 15mA. When the screen-grid voltage V_{g2} is equal to 170V a bias of $-3{\cdot}5$V is required for the control grid V_{g1}. The total cathode current I_k will then be 18mA, so for cathode bias a cathode resistor R_k of 195Ω would be needed. The nearest preferred value is 180Ω.

To maintain the full a.c. gain of the stage the cathode resistor must be adequately bypassed. The a.c. gain will then be greater than the d.c. gain, resulting in a considerable loss of the d.c. component of the video signal. The ratio of a.c. gain to d.c. gain in the example above, with the cathode resistor fully bypassed, would be about 4: 1. In some applications, however, the required a.c. gain can be achieved with only a part of the cathode resistor bypassed. In this way the a.c. gain and d.c. gain values can be maintained more nearly equal.

In a practical receiver circuit, shown in Fig. 11 (R_L = 2·7kΩ, R_{g2} = 4·7kΩ, R_k = 180Ω decoupled), the stage gain is more than 30 times.

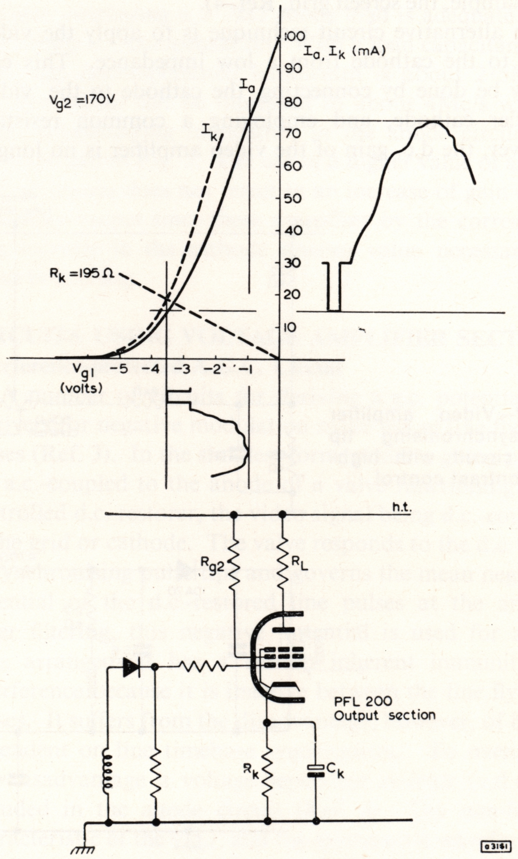


Fig. 10—Bias arrangements for positive modulation video amplifier

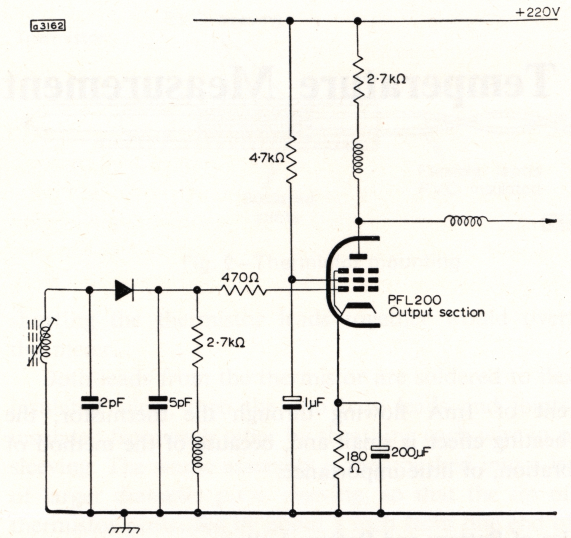


Fig. 11—Practical video amplifier circuit for positive modulation application

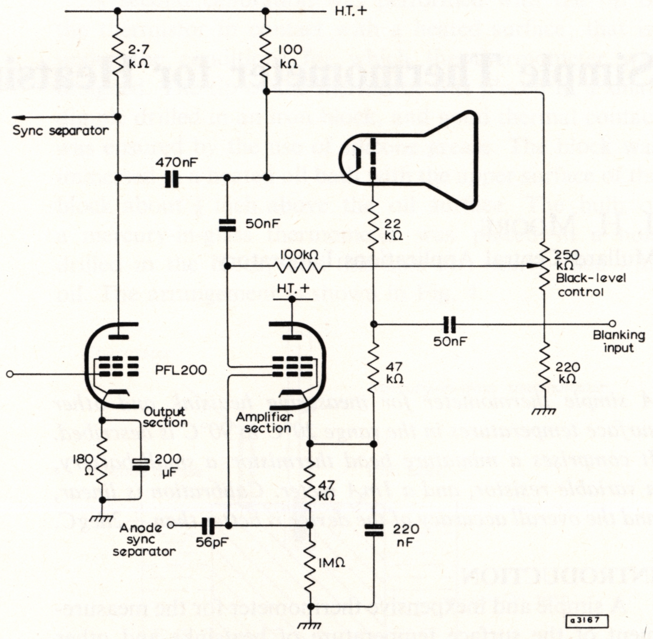


Fig. 12—A.C.-coupled black-level correction circuit

DUAL-STANDARD VIDEO AMPLIFIER CIRCUITS

The video amplifier in a dual-standard receiver has to operate with either a positively or negatively modulated vision signal and hence its operating bias requires switching. Alternatively the video signal from the detector can be a.c.-coupled to the amplifier. This alternative eliminates switching but the d.c. component of the signal is lost and the amplifier requires to operate in the class A model A.C.-coupling of a positively modulated signal and cause difficulty when interference pulses are considered and an alternative arrangement is to d.c.-couple the stage on positive modulation, and a.c.-couple with a d.c. restoring diode on negative modulation. The video stage then operates with a positive-going input signal on both standards and the synchronising tips are all at zero potential.

CIRCUITS USING VOLTAGE AMPLIFIER SECTION

The principal applications of the PFL200 amplifier section in circuits for positive modulation are as a synchronising pulse separator, which has been described on page 38. or as a black-level stabilising valve.

Black-level Stabilisation Circuits

Mean-level a.g.c. systems are widely used in receivers designed for positive modulation and in dual-standard receivers. A major shortcoming of the mean-level system is the effective suppression of the d.c. component of the video signal. This cannot be reinstated by a simple restoring diode as the amplitude of the synchronising pulses varies with the mean level of the video signal.

The d.c. component can, however, be reinstated in a displayed picture by a.c.-coupling the video signal to the picture tube cathode and supplying the d.c. compensating component to the control grid (Ref. 5). This technique is particularly advantageous in a dual-standard receiver and a suitable circuit employing the amplifier section for black-level stabilising is described below.

In the circuit shown in Fig. 12 the video signal is applied to the cathode of the picture tube with an a.c. coupling and also to the screen grid of the brightness control valve. The mean screen-grid potential is varied to form the brightness control, and a fixed d.c. level appropriate value is applied to the picture tube cathode in the normal way.

The valve is driven into conduction by the trailing edge of the separated synchronising pulse. The cathode capacitance is charged to a d.c. potential corresponding to the black-level of the video signal applied to the screen grid. This potential follows variations in the video signal black-level and is applied to the picture tube grid as the correction signal. The top of the brightness control potentiometer is connected to the picture tube cathode to improve the gain of the system.

The correction circuit is independent of the a.g.c system, and furthermore, the operation of a receiver is simplified since the brightness 'tracks' changes in the contrast control setting.

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(Subsequently published in **Wireless World,** Vol. 70, No. 5, May 1964, pp. 212 to 216.)

Simple Thermometer for Heatsink Temperature Measurement

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A simple thermometer for measuring heatsink and other surface temperatures in the range 20°C to 90°C is described. It comprises a miniature bead thermistor, a small battery, a variable resistor, and a 1mA meter. Calibration is linear, and the overall accuracy of the device is better than $\pm$2degC.

INTRODUCTION

A simple and inexpensive thermometer for the measurement of the surface temperature of heatsinks and other devices is described. The temperature range to be covered is from 20°C to 90°C, and the required accuracy is $\pm$2degC throughout this range.

The design of the thermometer has been achieved by using a very simple circuit which results in a linear calibration curve which can be easily obtained by determination of two or more fixed points (20°C, 90°C).

The factors governing the design, the details of the method of calibration, the errors, and the use of the thermometer, are described in the following sections.

FACTORS GOVERNING DESIGN

The circuit had to be as simple and as cheap as possible. This requirement suggested a series circuit consisting of a milliammeter, a battery, a thermistor, a variable resistor, and a 'push to operate' switch. The variable resistor is for calibration purposes. The switch is included to conserve the battery power by preventing the circuit from being inadvertently left on. The circuit is shown in Fig. 1.

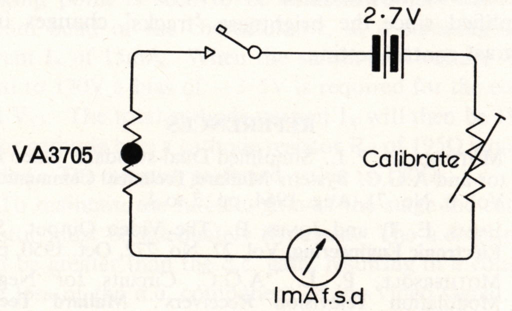

Fig. 1—Circuit of thermistor thermometer

Choice of Operating Current

As meters with a full-scale deflection of 1mA are reasonably cheap, the maximum current passed through the circuit was set to be 1mA, this current being passed when the thermistor temperature is 90°C. With a maximum current of 1mA flowing through the thermistor, the self-heating effect is small and, because of the method of calibration, of little importance.

Choice of Battery and Battery Voltage

The choice of battery is important, as the calibration of the circuit is very dependent on the battery voltage. Normal dry cells are ruled out, as they show a marked fall of voltage throughout discharge. On consideration, mercury cells seem to be an ideal alternative, as they provide a substantially constant voltage over most of their life. This choice limits the range of battery voltages to multiples of 1·35V. It was found that two mercury cells connected in series to give an e.m.f. of 2·7V resulted in the most linear calibration curve.

Choice of Thermistor

The choice of thermistor is determined by the following conditions:

(i) It should have a small thermal mass, so as not to affect appreciably the temperature of the surface to be measured.

(ii) It should have a quick response to external temperatures. This necessitates the mounting of the thermistor element in a good conductor of heat, which should preferably be an electrical insulator. Glass is a suitable material as it also provides mechanical protection of the thermistor element.

(iii) It should have a high temperature coefficient of resistance ('B' factor). This enables the change of resistance for a given change of temperature to be as high as possible, so that the change of current is as great as possible.

(iv) The resistance of the thermistor at 90°C must not exceed 2·7kΩ, otherwise – with a battery voltage of 2·7V – a full-scale deflection current of 1mA will not be attained.

A Mullard Varite thermistor type VA3705 was selected, as its resistance is 6·8kΩ $\pm$20% at 25°C, and about 700Ω at 90°C. Its 'B' factor is the highest in the range of thermistors with suitable resistance values. It is in miniature glass bead construction and is designed for rapid response to temperature changes.

Fig. 2 shows a simple way of mounting the thermistor and insulating its leads. This latter point is important, as shorting the thermistor leads together would overload the meter.

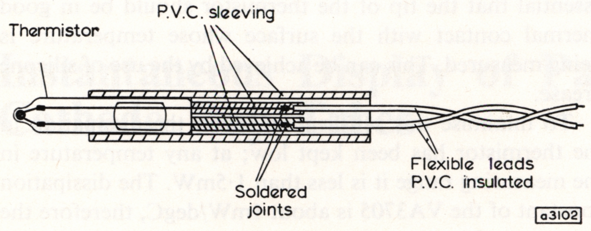

Fig. 2—Thermistor mounting

Both leads from the thermistor are soldered to flexible insulated wires, then the two bare leads and joints are insulated from each other by sheathing each wire in p.v.c. sleeving. The whole assembly is then pushed into a length of larger diameter p.v.c. sleeving, so that the tip of the thermistor protrudes by about ½ inch from one end of the sleeving. The other end of the sleeving extends well past the joints between the leads and the flexible conductor. The diameter of the outside sleeving is chosen so that it closely fits the glass encapsulation of the thermistor.

CALIBRATION

The variable resistor was set at its maximum value, and the thermistor was fully immersed in an oil bath which was at a temperature of 90°C. The variable resistor was then adjusted until the meter indicated full scale deflection. Meter readings were then taken with the oil temperature reduced by steps of 5degC, down to 20°C, the setting of the variable resistor being left unchanged. The calibration curve obtained is shown in Fig. 3.

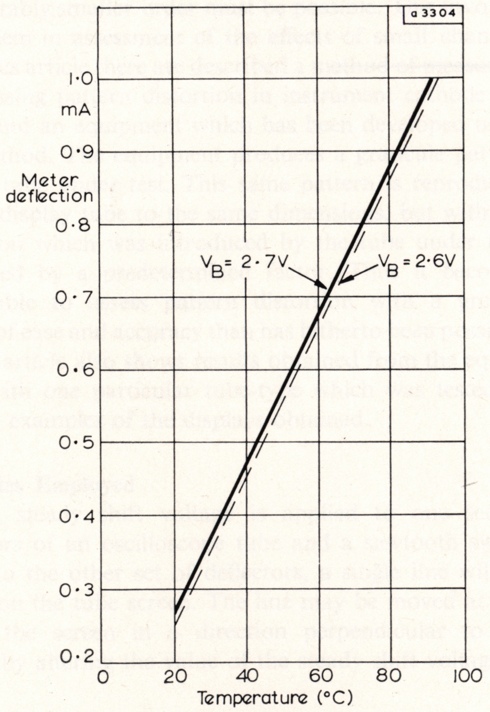

Fig. 3—Calibration curves obtained with one sample of VA3705 thermistor
—— with battery voltage of 2·7V
– – – with battery voltage of 2·6V

A second calibration was performed with the tip of the thermistor in contact with a heated surface; that is, under the conditions for which the thermometer was designed. The tip of the thermistor was placed in a small dimple drilled in an iron block, and good thermal contact was ensured by the use of silicone grease. The block was immersed in a heated oil bath with the upper surface of the block about $\frac{1}{8}$ inch above the oil surface. The bulb of a mercury-in-glass thermometer was placed in a hole drilled in the block, and this hole was nearly filled with oil. The arrangement is shown in Fig. 4.

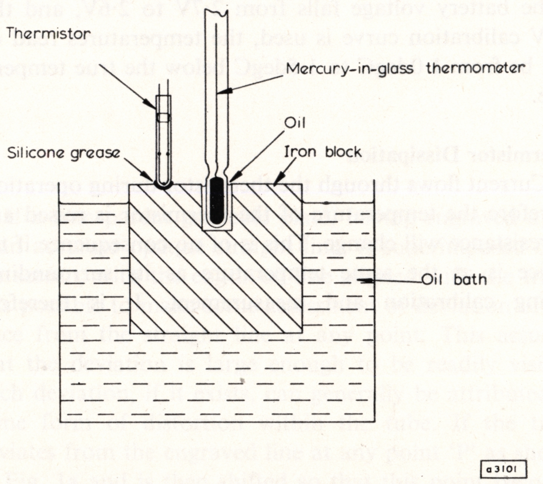

Fig. 4—Calibration method for surface temperature measurement

With the earlier resistor setting unchanged, meter readings were taken at 5degC intervals from 90°C down to 20°C, as indicated by the mercury-in-glass thermometer. The resulting calibration curve showed negligible deviation from the curve shown in Fig. 3.

ERRORS

Errors may be introduced by calibration discrepancies, variation of characteristics between one thermistor and another, battery voltage changes, and dissipation in the thermistor.

Calibration Method

If the circuit is calibrated with the thermistor fully immersed in oil, and is then used for surface temperature measurement with only the tip of the thermistor in contact with the surface, there may be some discrepancy between the true temperature of the surface and the temperature read off the calibration curve.

However, the check calibration that has been described (Fig. 4) has shown that such errors will be negligibly small. In practice, therefore, the circuit can be calibrated in one way (the simple immersion method) and used in another (the measurement of surface temperature) without this difference in conditions causing significant error.

Thermistor Spreads

Individual thermistors will not have precisely identical characteristics, therefore every circuit that is built must be separately calibrated. It is not acceptable to calibrate one circuit and then to use the same value of series resistance, and the same calibration curve, for all other circuits.

Battery Voltage

As this simple circuit is very dependent on the battery voltage, measurements have been made to determine the error incurred if the circuit is calibrated at 2·7V and is then used at a lower voltage without recalibration.

The broken line in Fig. 3 is the calibration curve obtained with a battery voltage of 2·6V. It shows that, if the battery voltage falls from 2·7V to 2·6V, and the 2·7V calibration curve is used, the temperatures read off will be from 1·0degC to 1·5degC below the true temperature.

Thermistor Dissipation

Current flows through the thermistor during operation, therefore the temperature of the thermistor is raised and its resistance will change. This is of no consequence if the device is at the same temperature as its surroundings during calibration and measurement. It is therefore essential that the tip of the thermistor should be in good thermal contact with the surface whose temperature is being measured. This can be achieved by the use of silicone grease.

To minimise this self-heating effect, the dissipation in the thermistor has been kept low; at any temperature in the measuring range it is less than 1·5mW. The dissipation constant of the VA3705 is about 1mW/degC, therefore the error caused by imperfect thermal contact will not exceed 2degC.

Overall Accuracy

If the circuit is calibrated with reasonable care, the battery voltage is correct, and good thermal contact is ensured during temperature measurement, an overall accuracy of better than ±2degC can be achieved.

Occasional recalibration is advisable, even if only as a reassurance. The drifts in the thermistor characteristic, in the resistor value, and in the accuracy of the milliammeter, will not be great. The most likely source of error is an unnoticed fall in battery voltage.

Instantaneous Display of Pattern Distortion in Cathode Ray Tubes

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Mullard Applications Research Laboratory

With the high degree of accuracy in c.r.t. displays there is an interest in the measurement of extremely small distortions introduced by the tube. Described in this article is an equipment which senses the distortion introduced by the tube. The distortion is magnified by a suitable factor and displayed on a second tube.

INTRODUCTION

Modern equipment using cathode ray tubes requires an ever increasing accuracy of display. As any distortions are frequently caused by more than one factor, it is desirable to be able to see the effect of extremely small changes in such distortions. In the applications of tubes it is essential that any distortion should be barely visible. Thus during the development of tubes, examination of a considerably smaller order must be possible. This involves a problem in assessment of the effects of small changes.

In this article there are described a method of measuring or assessing pattern distortion in instrument cathode ray tubes, and an equipment which has been developed using this method. The equipment produces a graticule pattern on the tube under test. This same pattern is reproduced on the display tube to the same dimensions, but with the distortion which was introduced by the tube under test, magnified by a predetermined factor. Thus it becomes practicable to assess pattern distortion with a greater degree of ease and accuracy than has hitherto been possible.

The article also shows results obtained from the equipment with one particular tube-type which was tested to provide examples of the displays obtained.

Principles Employed

If a steady shift voltage is applied to one set of deflectors of an oscilloscope tube and a sawtooth sweep signal to the other set of deflectors, a single line will be traced on the tube screen. The line may be moved at will across the screen in a direction perpendicular to the sweep, by altering the value of the steady shift voltage.

If a transparent graticule plate, having engraved upon it an accurate opaque straight line, is superimposed over the screen, with the engraved line aligned with the trace, it is possible to make some assessment of deviation of the trace from the straight line at any point. This assumes that the deviation is large enough to be readily visible. Such deviation, if it exists, can generally be attributed to some form of distortion within the tube. If the trace deviates from the engraved line at any point 'P' as shown in Fig. 1a and is then shifted so that this point becomes aligned with the engraved line as in Fig. 1b, then the change in shift voltage necessary to bring about this re-alignment is proportional to the magnitude of the deviation.

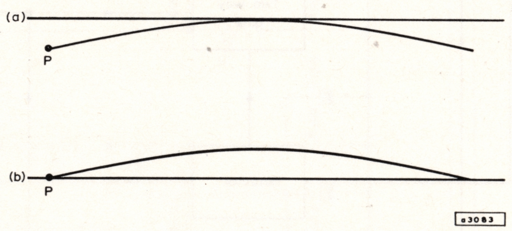

Fig. 1—Correction of error on trace

The use of some form of amplifying device to magnify the change in shift voltage by a factor of, say, 10 times, would result in a greatly enhanced indication of the deviation.

Practical Application of Principle

In the equipment described, the trace is aligned behind the engraved opaque line and a system is used whereby any deviation of the trace from the straight line is sensed by a photomultiplier. The output from the photomultiplier is converted, by means of a suitable detector system, into a shift signal. This shift signal is in turn amplified and applied to the tube deflectors in such a manner as to cause the trace to be re-aligned with the engraved line.

This article describes the principles of operation of an equipment for measuring small distortions in the display of cathode ray tubes. Circuits of only the main parts of the equipment are given, but interconnections are shown where relevant.

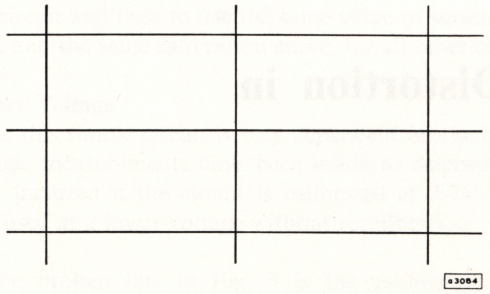

Fig. 2—Elementary lattice used on equipment

If the gain of the system is sufficiently high, the trace may be made to follow the line to a high degree of accuracy. The shift voltage which is applied will, at any time, accurately represent the deviation of the original uncorrected trace. If the shift voltage at the deflectors of the tube is amplified by a factor of exactly 10 times and is displayed on a second tube against the same sawtooth sweep as was applied to the tube under test, the resulting trace gives a picture with 10 times the original distortion.

Although it is of interest to view the distortion or deviation of a single trace, which may be moved at will over the tube screen, it is much more desirable to be able to view distortion over the screen generally. For this reason, in the practical equipment, six separate lines are traced sequentially, three equally spaced horizontal lines and three equally spaced vertical lines, as shown in Fig. 2.

Practical Circuits

The block diagram for the equipment is shown in Fig. 3. A master oscillator and locking strip are used to generate all of the frequencies and waveforms necessary for the timing of the gates and driving the deflection amplifiers. The appropriate waveforms from the master oscillator section of the circuit are combined in the graticule generator to give the waveforms necessary to produce the required display on the tubes. These signals are then amplified by x and y amplifier systems before application to the tube deflectors.

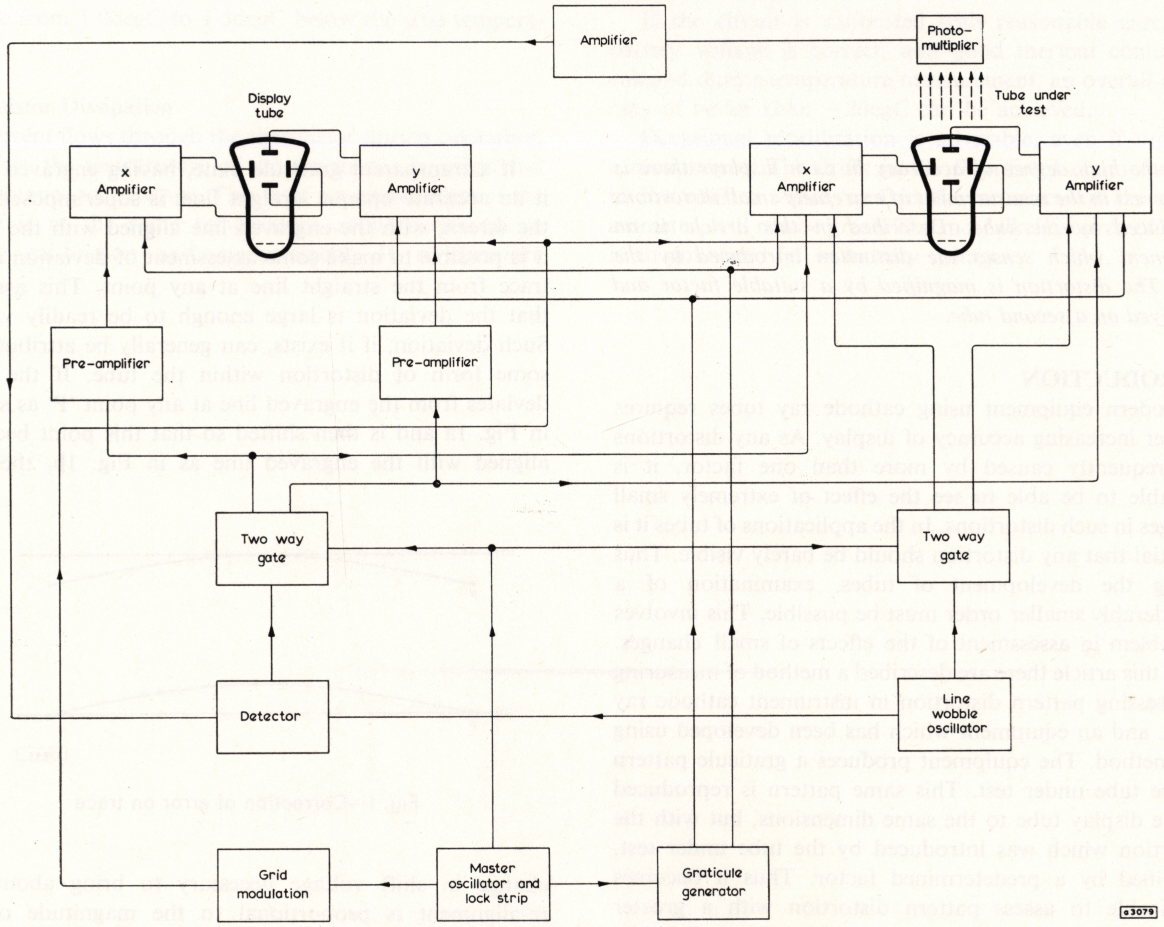
Amplifier
Photo-multiplier
Display tube
Tube under test
x Amplifier
y Amplifier
x Amplifier
y Amplifier
Pre-amplifier
Pre-amplifier
Two way gate
Two way gate
Detector
Line wobble oscillator
Grid modulation
Master oscillator and lock strip
Graticule generator
c3079

Fig. 3—Block diagram of equipment

The line wobble oscillator produces a high frequency sinewave signal for sensing purposes. This is switched alternately between the x and y axis, by means of a two way gate, in time with the graticule forming signals. The signal is superimposed on the graticule waveform and appears as an effective thickening of the line on the tube under test.

The tube is viewed by a photomultiplier, the output of which is amplified and fed to the detector. This detector

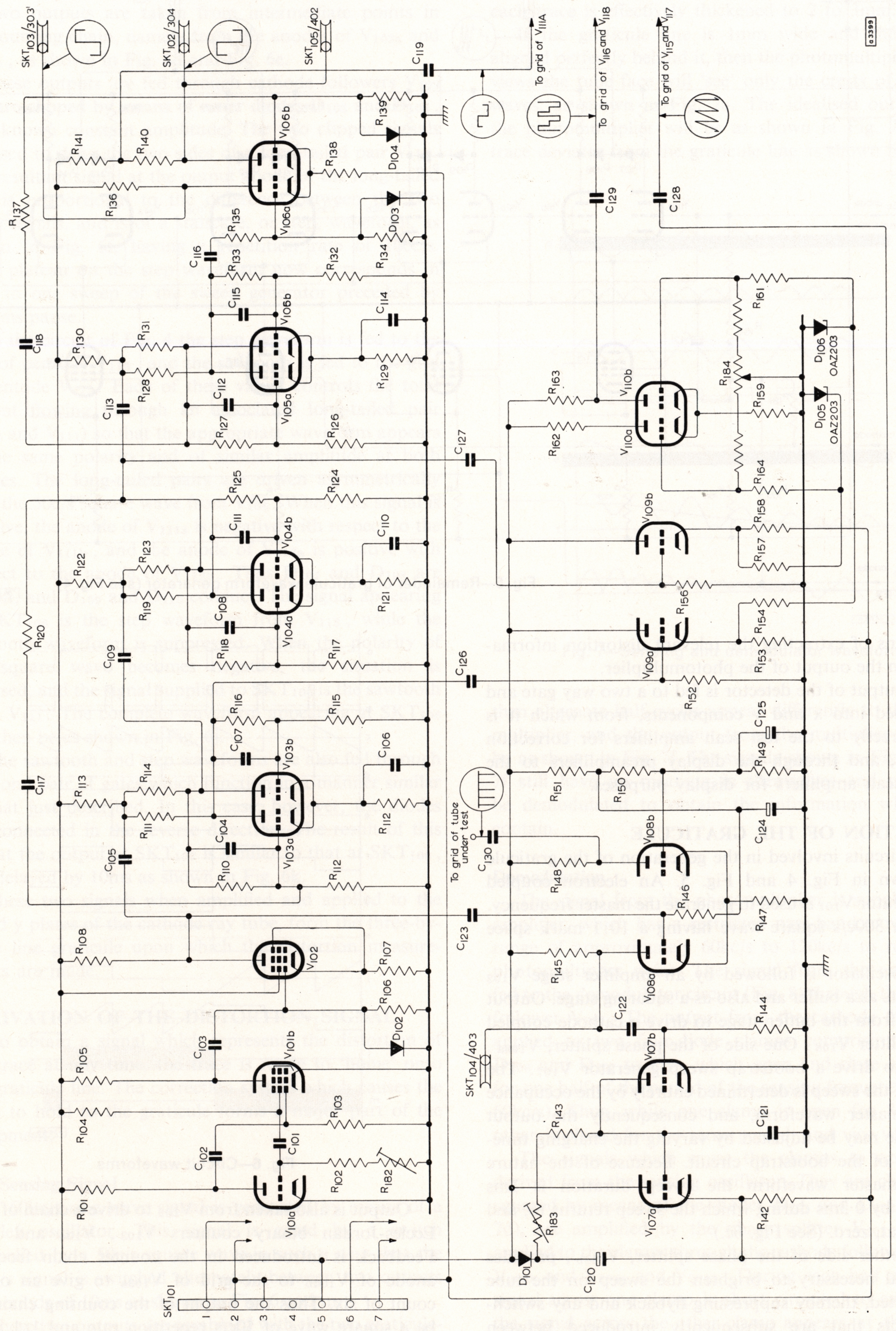
SKT103/303
SKT102/304
SKT105/402
To grid of V111A
To grid of V116 and V118
To grid of V115 and V117
To grid of tube under test
SKT104/403
SKT101
V101a
V101b
V102
V103a
V103b
V104a
V104b
V105a
V105b
V106a
V106b
V107a
V107b
V108a
V108b
V109a
V109b
V110a
V110b
D101
D102
D103
D104
D105 OAZ203
D106 OAZ203
c3399

Fig. 4—Part of graticule waveform generator (see Fig. 5)

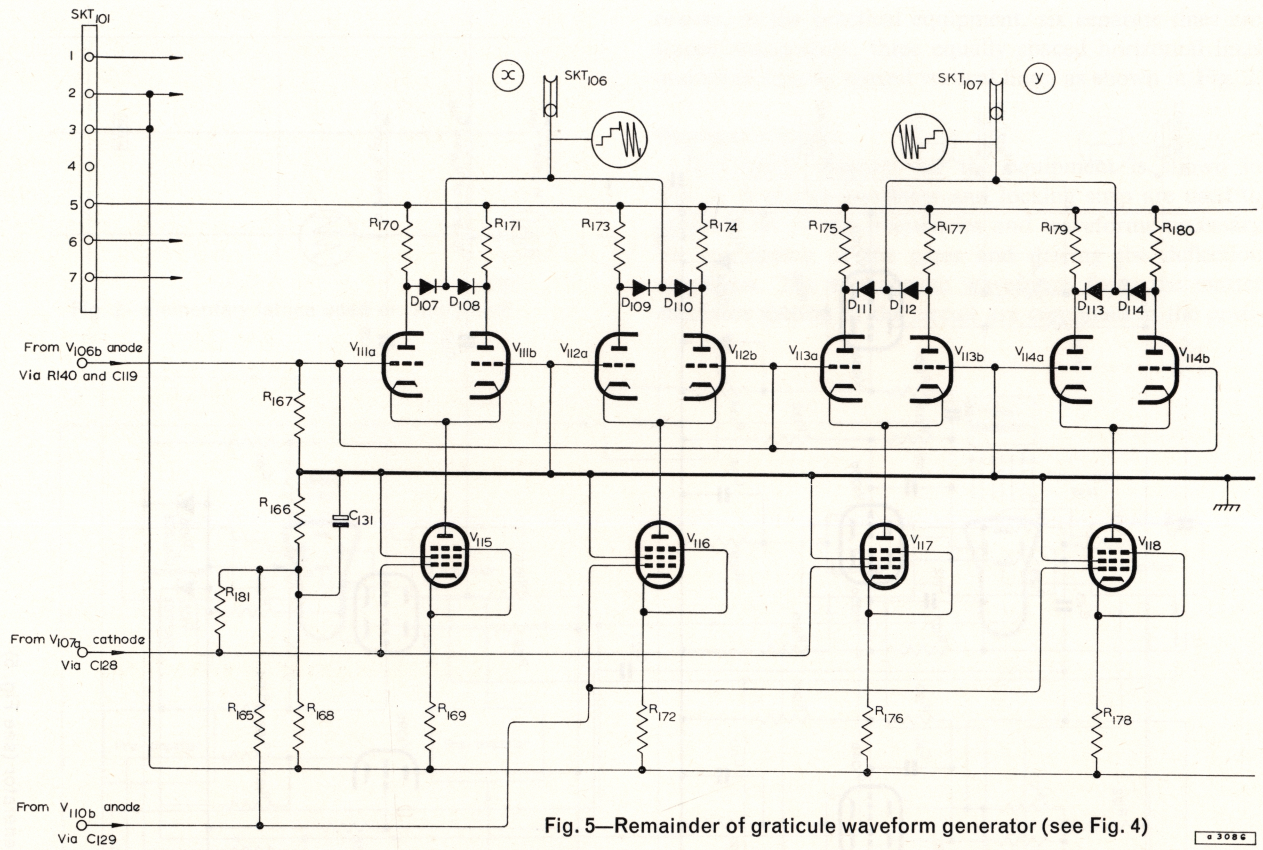

Fig. 5—Remainder of graticule waveform generator (see Fig. 4)

is a means of extracting the relevant distortion information from the output of the photomultiplier.

The output of the detector is fed to a two way gate and is switched into x and y components, from which it is taken directly to the test scan amplifiers for correction purposes, and through the display preamplifiers to the display scan amplifiers for display purposes.

FORMATION OF THE GRATICULE

The circuits involved in the generation of the graticule are shown in Fig. 4 and Fig. 5. An electron coupled multivibrator V_{101} is used to generate the master frequency. This is a 300c/s square wave having a 10:1 mark space ratio.

The oscillator is followed by an amplifier stage V_{102} which acts as a buffer and also as a squaring stage. Output is taken from the buffer stage to drive a cathode coupled phase splitter V_{108}. One side of the phase splitter, V_{108a}, is used to drive a bootstrap sweep generator V_{107}. The timing of the sweep is determined entirely by the occupance of the master waveform, and consequently the output amplitude may be adjusted by varying the charging time-constant of the bootstrap circuit. Because of the nature of the master waveform the sweep duration is 3ms followed by 0·3ms during which the sweep returns to, and remains at, zero. (See Fig. 6c.)

The other side of the phase splitter, V_{108b}, provides the signal necessary to brighten the sweep on the tube to be tested, thereby suppressing flyback and any switching effects that are subsequently introduced between successive sweeps.

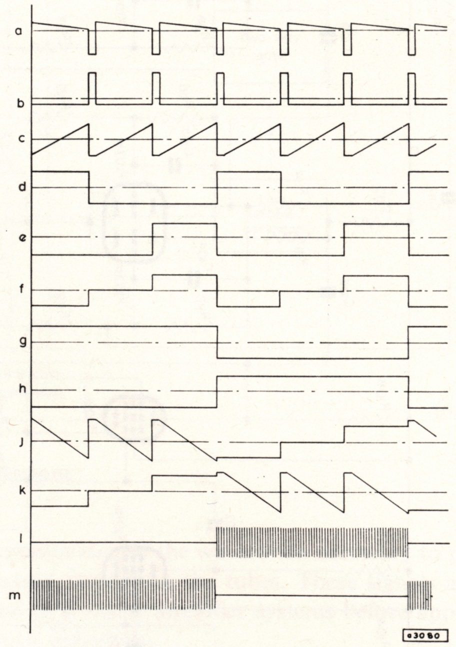

Fig. 6—Circuit waveforms

Output is also taken from V_{102} to drive a chain of three Eccles-Jordan binary counters V_{103}, V_{104} and V_{105}. Feedback is introduced in the counter chain from the anode of V_{104b} to the grid of V_{103a} to give an overall count of six. Thus the output of the counting chain will be a square wave of 50c/s repetition rate and 1:1 mark-space ratio. Output is taken symmetrically from V_{105} (that is, from both anodes), is clipped and squared, and is then used to drive a long-tailed pair, giving a balanced push-pull output for gating purposes (V_{106}).

Two outputs are taken from intermediate points in the counting chain, namely from the anodes of V_{103a} and V_{104b}, as shown in Fig. 6d and Fig. 6e.

These outputs are fed through cathode followers V_{109} and are clipped by means of zener diodes D_{105} and D_{106}, to a known constant amplitude. The two clipped signals are used to drive the two sides of a long-tailed pair V_{110}. The resultant signal at the output anode of the long-tailed pair is proportional to the difference between the two input signals, and is of a staircase, or step, waveform, as shown in Fig. 6f, having a repetition rate of 100c/s. Each plateau on the step waveform now corresponds in time to one sweep of the sweep generator preceded by a 0·3ms pause.

In the circuit of Fig. 4 the step waveform is fed to the grid of pentode V_{116}, and the sawtooth is fed to the grid of pentode V_{115}. Each of these valves controls the total current flowing through an associated long-tailed pair (V_{112} and V_{111}) so that the appropriate waveform appears in the same polarity and of similar amplitude at both anodes. The long-tailed pairs are driven asymmetrically with the 50c/s square wave from V_{106}. When this signal is positive, the anode of V_{111a} is negative with respect to the anode of V_{111b}, and the anode of V_{112a} is positive with respect to the anode of V_{112b}. Thus D_{107} and D_{108} are cut off, and D_{109} and D_{110} conduct. The signal appearing at SKT_{106} is the step waveform from V_{112}, while the sawtooth waveform is suppressed. When the polarity of the square wave becomes negative, the situation is reversed, and the signal supplied to SKT_{106} is the sawtooth from V_{111}. The complete waveform appearing at SKT_{106} will then be as shown in Fig. 6j.

The sawtooth and step waveforms are also fed through a second pair of gates which function in a manner similar to that just described. In this case, however, the diodes are connected in the reverse direction. The result of this is that the output at SKT_{107} is similar to that at SKT_{106}, but delayed by 10ms as shown in Fig. 6k.

These two signals when amplified and applied to the x and y plates of the cathode ray tube, form the three-by-three line graticule upon which the distortion measurements are made.

DERIVATION OF THE DISTORTION SIGNAL

To obtain a signal which represents the distortion of the trace at any time, the trace is made to 'home' onto the graticule line. The correction system which causes the trace to hold to the graticule forms a major part of the equipment.

The Sensing Signal

A 60kc/s sine wave signal is generated by means of a Hartley oscillator. This signal is gated by a switch controlled by a 50c/s square wave in such a way that the 60kc/s signal is switched alternately between two outputs for periods of 10ms as shown in Fig. 6 and m.

These two outputs are combined with the two graticule-forming signals applied to the tube under test. The timing of the signals is such that the 60kc/s signal is displayed on the sweep, that is, superimposed upon the step waveform. The amplitude of the sinewave is adjusted so that each trace is effectively thickened to 2 to 3mm.

If the graticule line is 1mm wide and the trace is aligned perfectly behind it, then the photomultiplier which views the tube face will 'see' only the crests of the sinewaves, as shown in Fig. 7a. The idealised output from the photomultiplier will be as shown in Fig. 7b. If the trace deviates from the graticule line as shown in Fig. 7e, then alternate half-cycles appear differently to the photomultiplier, and the output from the multiplier tube will be as shown in Fig. 7f. The signals shown in Figs. 7b and 7f still have no meaning in themselves and require to be demodulated to obtain the information which they contain.

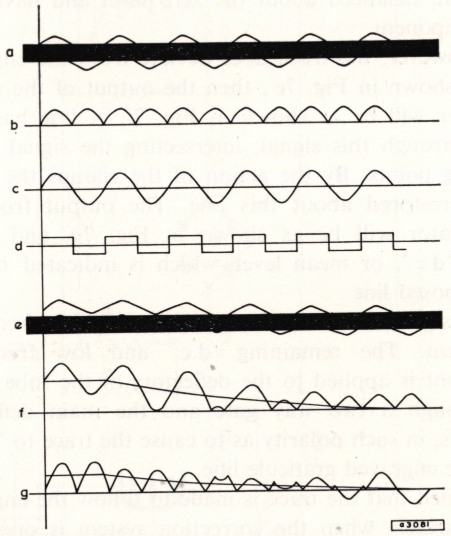

Fig. 7—Sensing and demodulator waveforms

Demodulation

The signal from the photomultiplier is supplied to its amplifier, which has a restricted pass-band, covering the range of approximately 60kc/s to 120kc/s to reduce the photomultiplier noise. The signal from the amplifier is fed to the demodulator circuit (Fig. 8) through the cathode follower V_{202}. The output from the cathode follower is applied to two four-diode clamps comprising D_{203} to D_{206} and D_{207} to D_{210} which open and close alternately for one half of the period of the sensing frequency (that is, 60kc/s). Thus, while one clamp is open for 8·33µs the other is closed for the same period, and vice versa.

The signals which cause the clamps to operate are derived from the sensing oscillator. They are clipped and squared by means of the diodes D_{201} and D_{202} (see Fig. 7d), are amplified by the phase splitter V_{201}, and are applied to the clamps through the cathode followers V_{203a}, V_{203b} and V_{204a} and V_{204b}. During each half period the signal across one clamp is short circuited to earth while the signal across the other clamp appears at the appropriate grid of V_{205}.

The signal applied to the grid of V_{205a} appears at the anode of V_{205b} without phase inversion, while that applied to the grid of V_{205b} is inverted at the same anode. Thus the overall signal appearing at the anode of V_{205b} contains the information provided by the photomultiplier but is inverted in polarity for alternate periods of 180° (or 8·33µs).

If the trace is correctly aligned behind the graticule line as in Fig. 7a, the signal from the photomultiplier will be as shown in Fig. 7b. After demodulation the signal appears as in Fig. 7c, that is, as a symmetrical waveform, balanced about the zero level and having no d.c. component.

If, however, the trace does deviate from the engraved line, as shown in Fig. 7e , then the output of the photomultiplier will be as shown in Fig. 7f. A line has been drawn through this signal, intersecting the signal at the switching points. By the action of the clamps the signal will be restored about this line. The output from the demodulator will be as shown in Fig. 7g, and has a definite 'd.c.', or mean level, which is indicated by the superimposed line.

This signal is filtered to remove the high frequency component. The remaining 'd.c.' and low frequency component is applied to the deflectors of the tube under test through a two way gate and the main deflection amplifiers, in such polarity as to cause the trace to 'home' on to the engraved graticule line.

Provided that the trace is made to follow the engraved line accurately when the correction system is operative, then the correction signal will be representative of the deviation of the trace at any given time. The higher the gain within the correction loop, the more accurate will be the following of the line, and the more accurate will be the representation. In the equipment described the error is small enough to be considered negligible.

Distortion Signal Gates

The distortion signal obtained from the demodulator is a continuous signal from only one output. It is therefore necessary to gate the output from the demodulator to obtain distinct and separate x and y correction signals.

The output from the demodulator is fed to two four-diode gates D_{301} to D_{304} and D_{305} to D_{308} (shown in Fig. 9) which are opened and closed alternately by means of the 50c/s square waveforms obtained from V_{106}. During the period in which the horizontal or x lines are being traced, the gate D_{301} to D_{304} closes and the grid of V_{301b} is short circuited to earth. The gate D_{305} to D_{308} is open and allows signal to pass to the grid of V_{305a}. This means that correction is applied to the y deflectors of the tube. While vertical or y lines are being traced, the situation is reversed and a correction signal is applied via the grid of V_{301b} to the x deflectors of the tube.

Although the four-diode gate is fundamentally a balanced device it has been found that it is still possible to get some switching break-through which appears as a pedestal to the distortion signal. If the switching action is clean this pedestal will not introduce any significant error into the system in terms of distortion but can nevertheless bring about a slight constant displacement

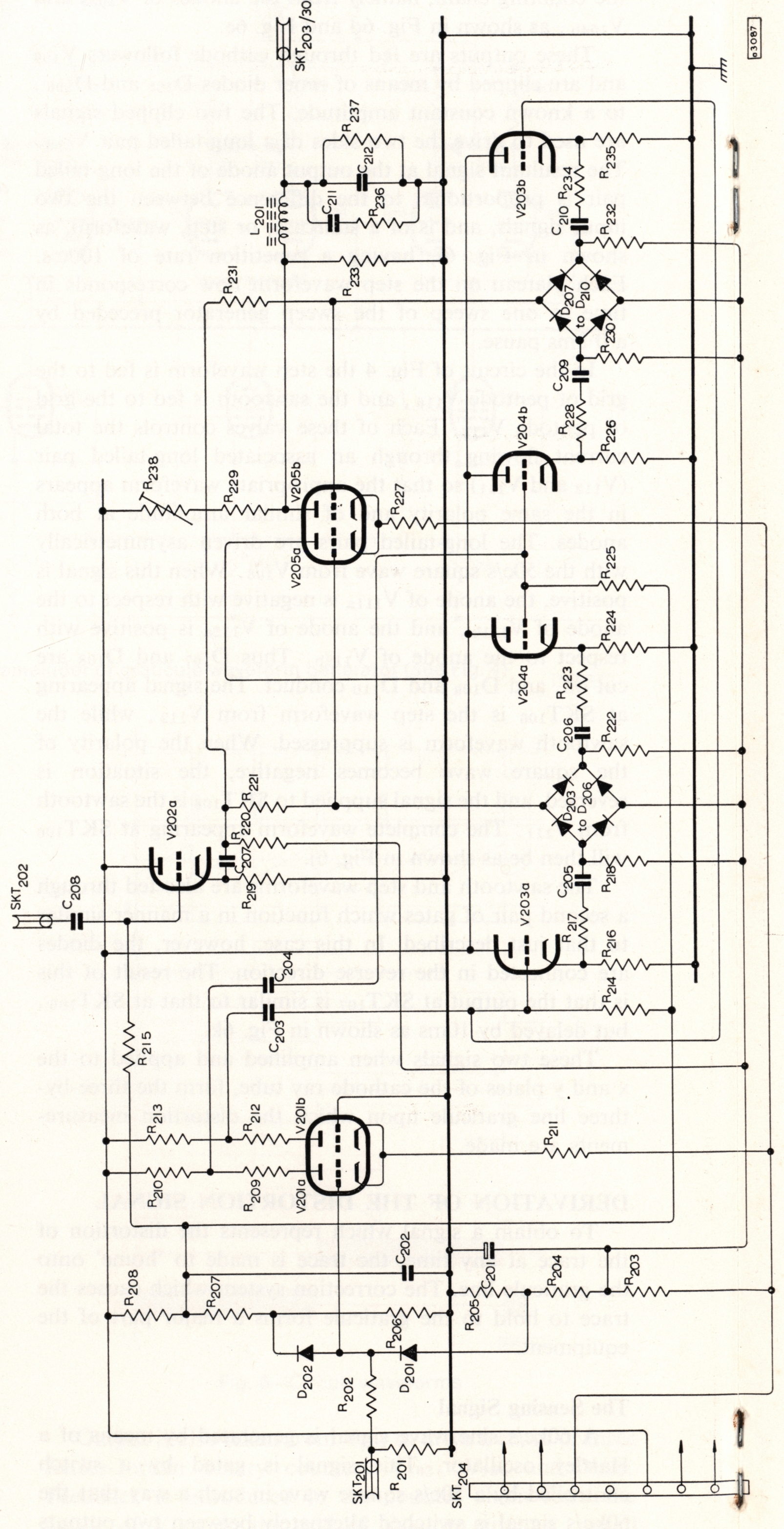

Fig. 8—Demodulator circuit

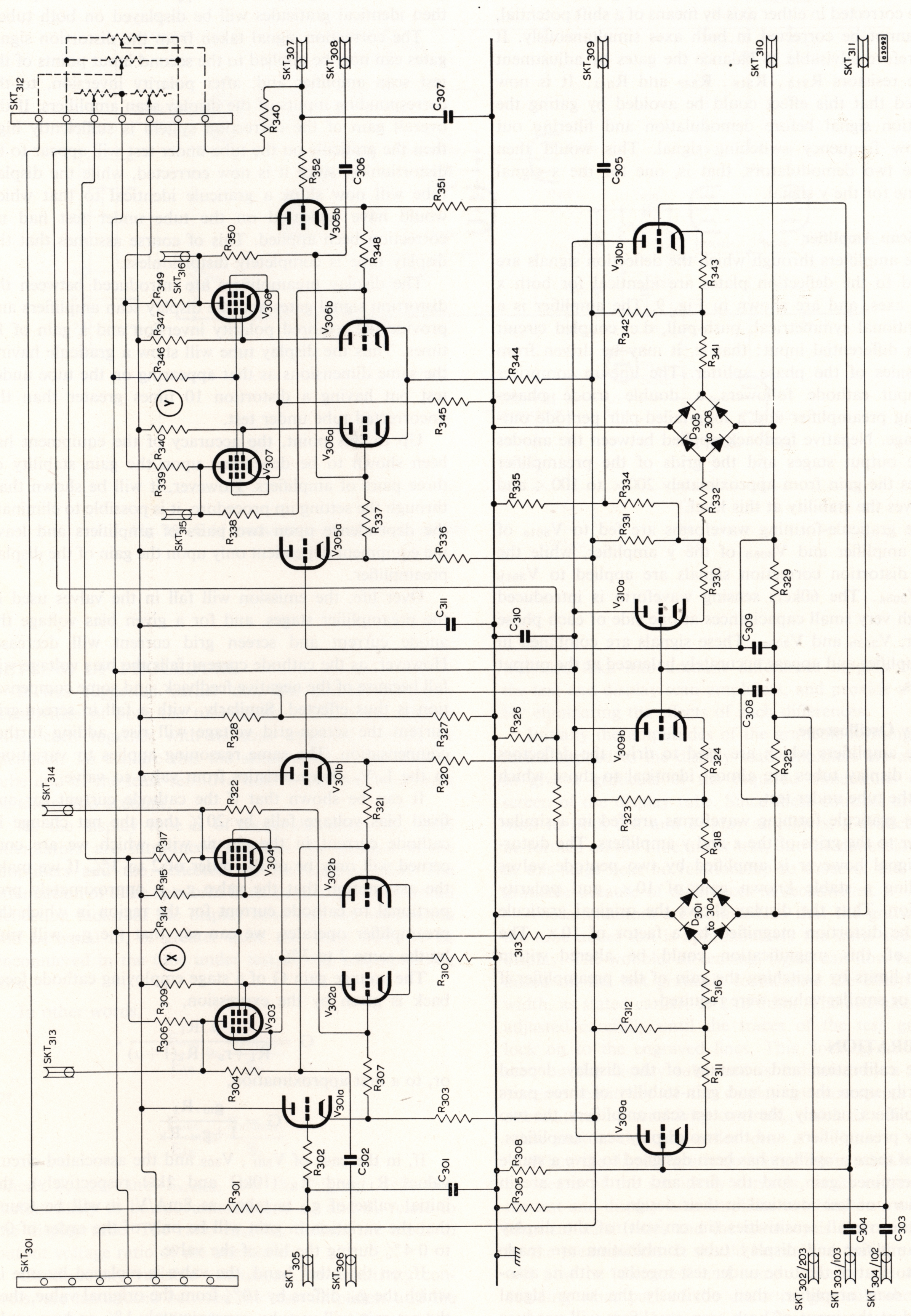

Fig. 9—Test scan amplifiers and distortion gates

of the trace with respect to the engraved graticule. This can be corrected in either axis by means of a shift potential, but cannot be corrected in both axes simultaneously. It is therefore advisable to balance the gates by adjustment of the resistors R_{316}, R_{318}, R_{332} and R_{341}. It is now believed that this effect could be avoided by gating the distortion signal before demodulation and filtering out the low frequency switching signal. This would then require two demodulators, that is, one for the x signal and one for the y signal.

Test Scan Amplifier

The amplifiers through which the deflection signals are applied to the deflection plates are identical for both x and y axes, and are shown in Fig. 9. The amplifier is a conventional symmetrical, push-pull, d.c. coupled circuit having differential input; that is, it may be driven from both sides of the phase splitter. The line-up comprises the input cathode followers, a double triode phase-splitting preamplifier and a long-tailed-pair pentode output stage. Negative feedback applied between the anodes of the output stages and the grids of the preamplifier reduces the gain from approximately 200× to 100× and improves the stability at this level.

The graticule-forming waveforms are fed to V_{301a} of the x amplifier and V_{305b} of the y amplifier, while the gated distortion correction signals are applied to V_{301b} and V_{305a}. The 60kc/s sensing waveform is introduced through very small capacitances at one side of each phase splitter, V_{302b} and V_{306a}. These signals are combined in the amplifier and appear accurately balanced at the output anodes.

Display Oscilloscope

The amplifiers which are used to drive the deflectors of the display tubes are almost identical to those which drive the tube under test.

The graticule forming waveforms are fed in a similar manner to the grids of the x and y amplifiers. The distortion signal however is amplified by two pentode valves providing a stable known gain of 10×, and polarity inversion. Thus the display shows the original graticule with the distortion magnified by a factor of 10×. The value of this magnification could be altered within certain limits by switching the gain of the preamplifier if larger or smaller values were required.

CALIBRATION

The calibration and accuracy of the display depend primarily upon the gain and gain-stability of three pairs of amplifiers, namely, the two test scan amplifiers, the two display preamplifiers, and the two display scan amplifiers. Each of these amplifiers has been designed to give a stable predetermined gain, and the first and third pairs are in fact more or less identical in their design.

If the overall sensitivities (in cm/volt) of the display scan amplifier and display tube combination are made equal to that of the tube under test together with its associated scan amplifier, then obviously the same signal applied at the input of both scan amplifiers will produce identical deflections on the two tubes. Thus if common graticule forming signals are applied to both amplifiers then identical graticules will be displayed on both tubes.

The correction signal taken from the distortion signal gates can now be applied to the second input points of the test scan amplifier and, after polarity inversion, to the corresponding inputs of the display scan amplifiers. If the overall gain of the correction system is sufficiently high then the graticule on the tube under test will appear to be distortionless since it is now corrected, while the display tube will now show a graticule identical to that which would have appeared on the tube under test had no correction been applied. This of course assumes that the display tube is completely distortionless.

The display preamplifiers are introduced between the distortion signal gates and the display scan amplifiers and provide the required polarity inversion and a gain of 10 times. Thus the display tube will show a graticule having the same dimensions as that appearing on the tube under test but having a distortion 10 times greater than the uncorrected tube under test.

Up to this point, the accuracy of the equipment has been shown to be dependent upon the gain stability of three pairs of amplifiers. However, it will be shown that, through the setting up procedure, it is possible to eliminate the dependence upon two pairs of amplifiers and leave the equipment dependent only upon the gain of the display preamplifier.

Over life, the emission will fall in the valves used in the preamplifier stages, and for a given bias voltage the anode current and screen grid current will decrease. However, as the cathode current falls the bias voltage will fall because of the negative feedback, and some compensation is thus effected. Similarly, with a fall in screen-grid current the screen-grid voltage will rise, adding further compensation. The same reasoning applies to variations in the I_a/V_g characteristics from valve to valve.

It can be shown that if the cathode current for any fixed bias voltage falls by 20% then the net change in cathode current in the circuit with which we are concerned will only be of the order of 2 to 3%. If we make the assumption that the valve g_m is approximately proportional to cathode current for the region in which the preamplifier operates, we can say that the g_m will vary by the same 2 to 3%.

The voltage gain G of a stage employing cathode feedback is given by the expression,

$$G = \frac{\mu . R_1}{R_1 + r_a + R_k(1+\mu)},$$

or, to a first approximation,

$$G = \frac{g_m . R_1}{1 + g_m . R_k}.$$

If, in the case of V_{501}, V_{502} and the associated circuit values R_1 and R_k (10kΩ and 1kΩ respectively), the initial value of g_m is taken as 8mA/V, it will be found that the variation in gain will be only of the order of 0·3 to 0·4% during the life of the valve.

If, on the other hand, the valve is replaced by one in which the g_m differs by 10% from the original value, then the net gain will vary by approximately 1%, and we could safely take ±2% as the limits of variation.

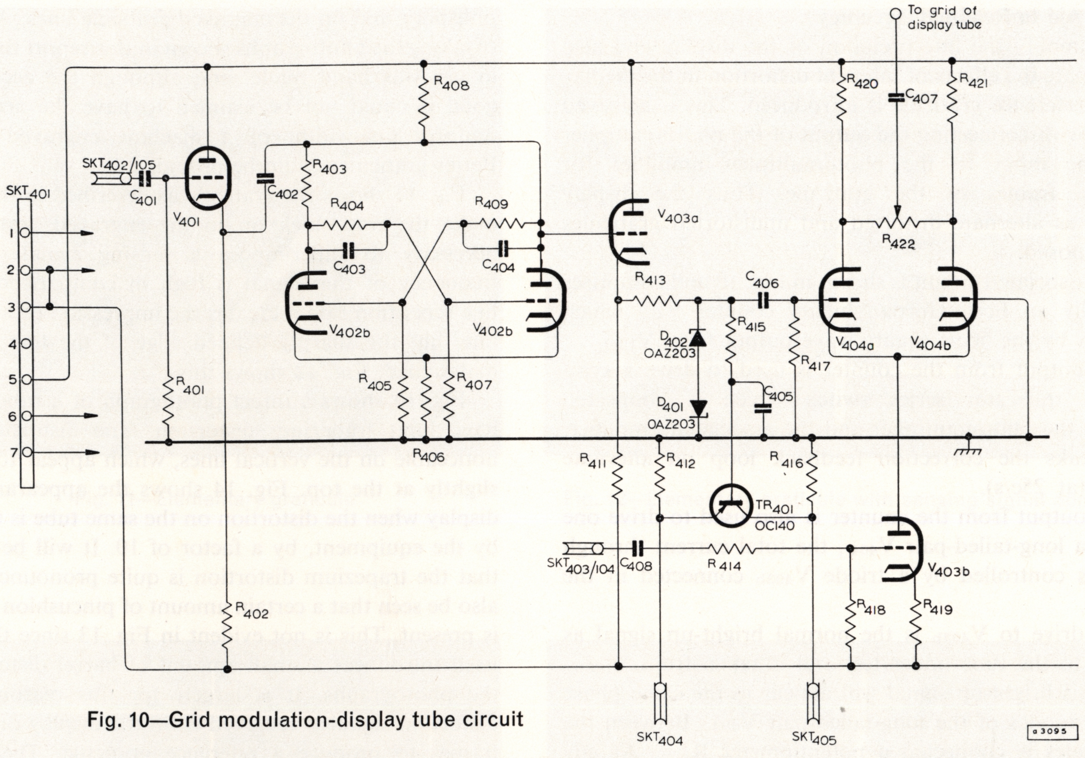

Fig. 10—Grid modulation-display tube circuit

SETTING UP PROCEDURE

It was suggested earlier that the overall deflection sensitivities of both the tube under test and the display tube, each together with its scan amplifier, should be identical. To achieve this the sensitivities of the display tube and of the least sensitive tube to be measured must be taken into account.

As the display tube is of moderate sensitivity, fixed attenuators are introduced between the outputs of the amplifiers and the deflectors of the display tube. The attenuation of these networks must be such that the overall deflection factor of the display tube and the attenuators will be equal to, or greater than, the worst likely to be encountered in the tube under test, in the corresponding axis.

In other words,

$$\frac{E_{out}}{E_{in}} \geqslant \frac{S_{y(max)}}{S_y}$$

or

$$\frac{E_{out}}{E_{in}} \geqslant \frac{S_{x(max)}}{S_x}$$

where $S_{y(max)}$ and $S_{x(max)}$ are the highest deflection factors likely to be encountered, and S_y and S_x relate to the display tube. The ratio E_{out}/E_{in} defines the input to output voltage ratio of the attenuator.

Continuously variable attenuators, or sensitivity controls, are introduced between the scan amplifiers and the tube under test, enabling the two tube-amplifier combinations to be matched for sensitivity. These attenuators will also take into account differences in gain between the test and display scan amplifiers, and provide a means for eliminating the effects of such differences.

Initially the amplitudes of the graticule-forming waveforms applied to the display scan amplifiers are adjusted to give a graticule of the required dimensions on the screen of the display tube. Since these controls are common to both test and display amplifiers, a graticule will be formed on the tube under test, but the dimensions may at this stage bear no relationship to those of the display graticule because of differences in sensitivities.

The attenuators are adjusted so that the graticule on the tube under test is approximately the same size as the display. The wobble, or sensing, signal can now be introduced on the test graticule to increase the effective line width, as stated earlier, to 2 to 3mm. The attenuators are adjusted carefully until the traces of the test graticule lock on to the engraved lines. This 'locking-on' can be seen on the display tube and is indicated by a sudden extreme sensitivity of the display to movements of the attenuators, or sensitivity controls, which at other times relate only to the tube under test.

When the display graticule is once again adjusted, now by means of the sensitivity controls, to its original size, the test and display tube-amplifier combinations will have identical sensitivities and the accuracy is determined only by the preamplifiers. It should be realised that the display will now be 'distorted' and some difficulty can arise in deciding what part of the distorted graticule should be set to the original dimensions. The errors arising from this difficulty, however, may be considered to be negligible.

Further Aid to Read-out Accuracy

To simplify the interpretation of the display and also to eliminate the effects of inherent distortion in the display tube, a reference graticule is introduced. This is achieved simply by disconnecting the output of the photomultiplier from the input to the photomultiplier amplifier for alternate frames of the graticule. Thus the display appears as alternate distorted and undistorted graticules superimposed.

The associated circuit is shown in Fig. 10 and comprises essentially an Eccles-Jordan binary counter V_{402} which is driven by the 50c/s square wave output from V_{106b}.

The output from the counter is used to drive a symmetrical transistor series switch which is connected between the photomultiplier and the associated amplifier, and breaks the correction feedback loop on alternate frames (at 25c/s).

The output from the counter is also used to drive one side of a long-tailed-pair V_{404}, the total current through which is controlled by a triode V_{403b} connected in the cathode.

The drive to V_{403b} is the normal bright-up signal as applied to the tube under test and is taken from V_{108b}. Thus, this bright-up signal will appear in the same phase at both anodes of the long-tailed-pair V_{404}. Between the two anodes is connected a potentiometer R_{422}. At any point on this potentiometer the normal bright-up will appear at constant amplitude; however, the 25c/s square wave produced by V_{402} will appear in opposite polarity at either end, and at zero amplitude at the centre.

Thus it is possible, without affecting the normal flyback suppression, to control, by means of R_{422}, the relative brightness of the undistorted graticule and the graticule showing the distortion.

The final display thus indicates a graticule with a superimposed graticule carrying an enlarged version of the distortion. The magnitude and direction of distortion can be seen at a glance and can also be measured. The distance between reference line and distorted line represents, in this case, ten times the true distortion. Typical displays are discussed below.

TYPICAL DISPLAYS

Below are a series of photographs taken from the equipment described. An explanation of each photograph is given.

General Comments

The tube type used for all of the photographs was the D13-22GH, because of its versatility in terms of pattern correction and manipulation of pattern shape.

This display exhibits a certain amount of ripple on the lines in each case. It was thought initially that this had its origin in noise from the photomultiplier, but it was noticed that the irregularities appeared to be stable and repeated on each frame. This led to the belief that they were caused by phosphor grain and small non-uniformities both in the phosphor and on the engraved graticule lines. At the time of taking the photographs shown in this report the graticule in use was hand made, and although the accuracy was good it could not be claimed to have the accuracy, or uniform line-width, of a machine engraved graticule. Better graticules could be provided.

Fig. 11 shows the graticule as described in the text. To make the trace lock on to the engraved graticule it is necessary to superimpose a sensing signal. Since the frequency of this signal is high in comparison with the line repetition rate it creates the impression of a thickened line, but the sharply defined edge of the line should be maintained. Fig. 12 shows this.

Fig. 13 shows a direct photograph of a tube found to have limit trapezium distortion. The distortion is just noticeable on the vertical lines, which appear to converge slightly at the top. Fig. 14 shows the appearance of the display when the distortion on the same tube is magnified, by the equipment, by a factor of 10. It will be seen now that the trapezium distortion is quite pronounced. It will also be seen that a certain amount of pincushion distortion is present. This is not evident in Fig. 13 since the camera itself introduces a small amount of barrel distortion into the photographs. It is largely for this reason that the distortion signal is removed from the display on alternate frames to produce a reference graticule. The reference graticule may be seen faintly in Fig. 14.

Fig. 15 shows the display obtained from a typical tube type D13-22GH with no correction potentials applied. A small amount of barrel distortion is present. Fig. 16 shows the tube mentioned in conjunction with Fig. 15 but with distortion correction applied. A minute amount of trapezium distortion remains on the outer horizontal lines and there is a slight curving of the outer vertical lines.

To correct for certain distortions (for example, barrel and pincushion), a potential is applied to the p.d.s. and p.d.s. skirt. This photograph (Fig. 17) shows the effect of applying excessive positive voltage with respect to y plate mean potential. It will be seen that barrel distortion is introduced. Fig. 18 shows the effect of excessive negative correction potential. It will now be seen that pincushion distortion has been introduced. Obviously a condition exists between Figs. 17 and 18 which would give an optimum correction, although ideal correction for horizontal and vertical lines may not occur simultaneously.

Fig. 19 shows the effect of increasing the p.d.a. ratio beyond the maximum quoted value. The correct ratio for the tube type used should be 10:1, but a ratio of 15:1 tends to increase the amount of barrel distortion present. Distortion introduced in this manner may generally be corrected by the means mentioned previously.

Fig. 20 shows the effect of rotation of the tube under test with respect to the engraved graticule. The actual rotation of the tube was approximately 1° and consequently this picture helps to give some appreciation of the sensitivity of the equipment to distortion.

It can be seen from the above examples that the equipment described provides a satisfactory means of viewing distortions of an order less than those discernible by eye, and provides a useful development and quality control instrument.

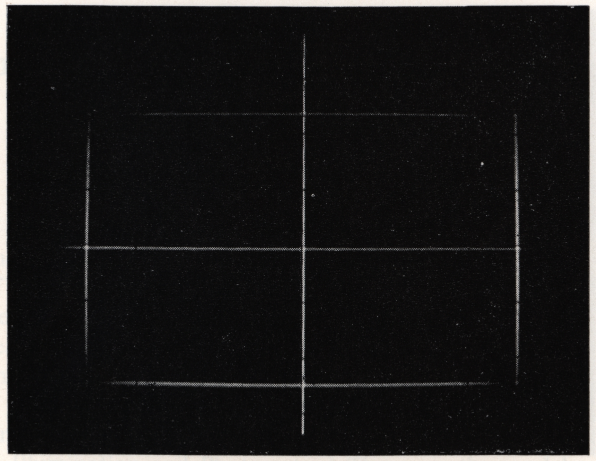

Fig. 11—Elementary graticule

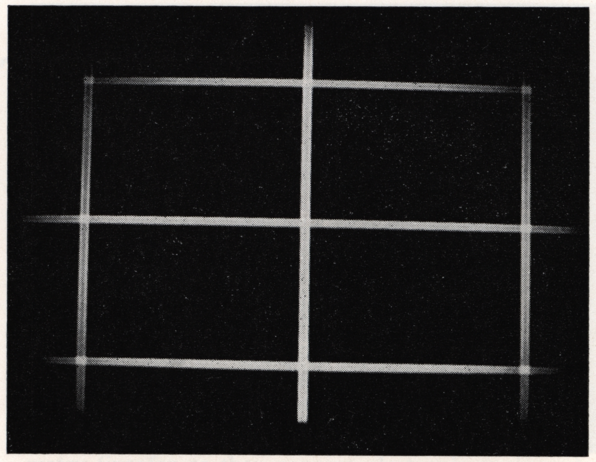

Fig. 12—Elementary graticule with sensing signal applied

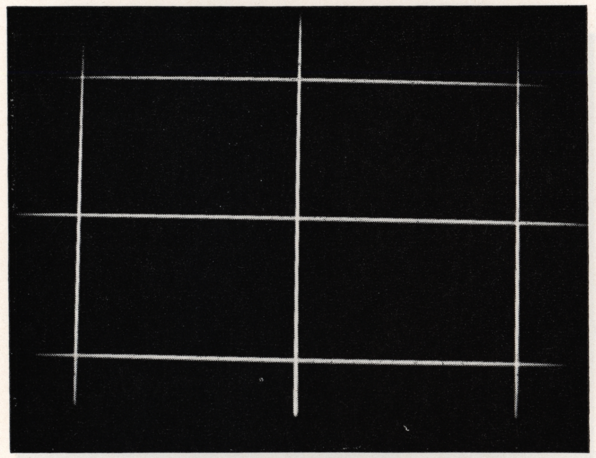

Fig. 13—Limit trapezium distortion in the y-axis

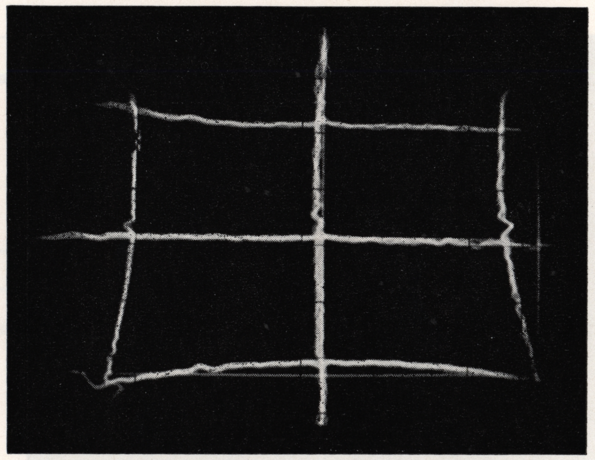

Fig. 14—Limit trapezium distortion with 10× magnification

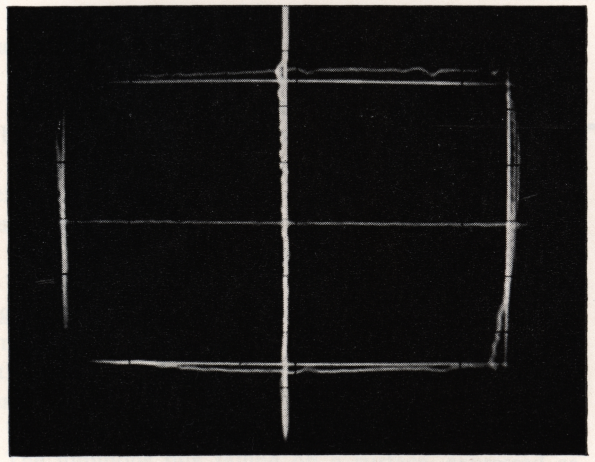

Fig. 15—Typical display with tube type D13-22GH

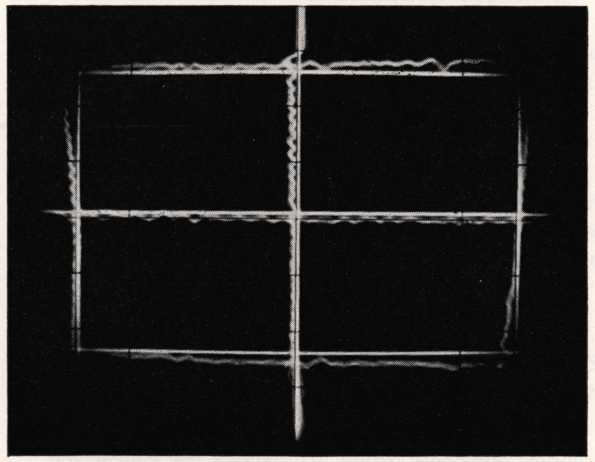

Fig. 16—Typical display with tube type D13-22GH with optimum correction applied

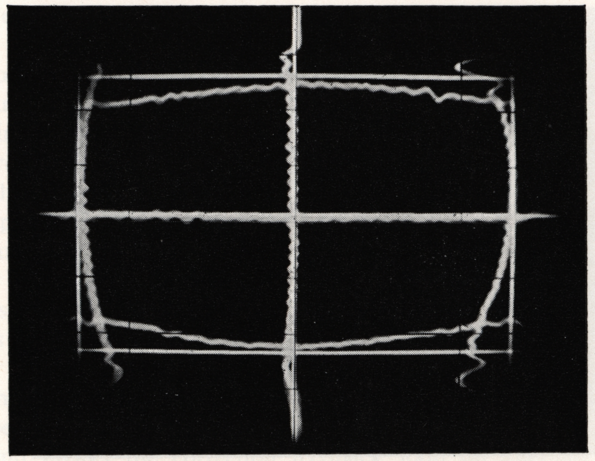

Fig. 17—Barrel distortion caused by +50V applied to mesh skirt

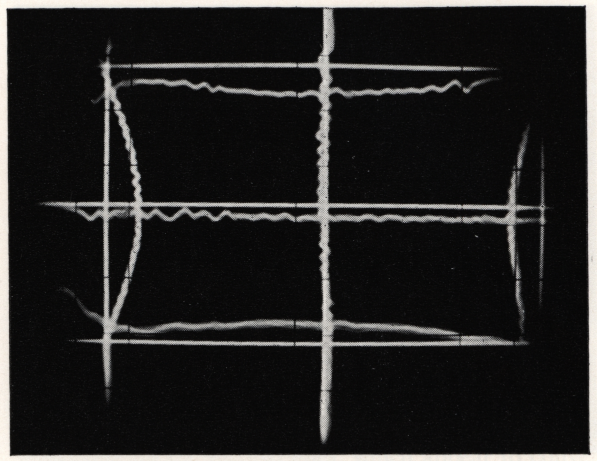

Fig. 18—Pincushion distortion caused by −50V applied to mesh skirt

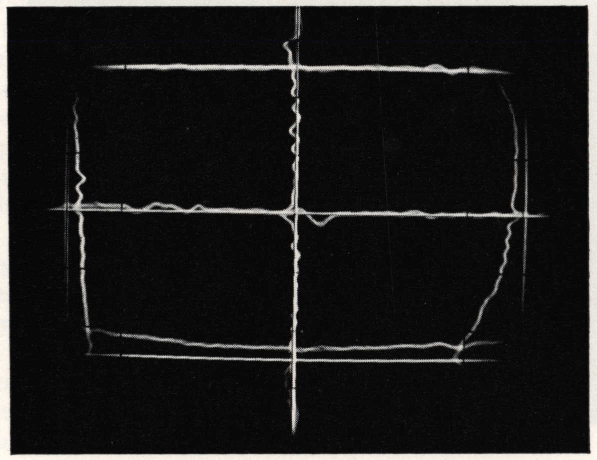

Fig. 19—Effect of excessively high p.d.a. ratio

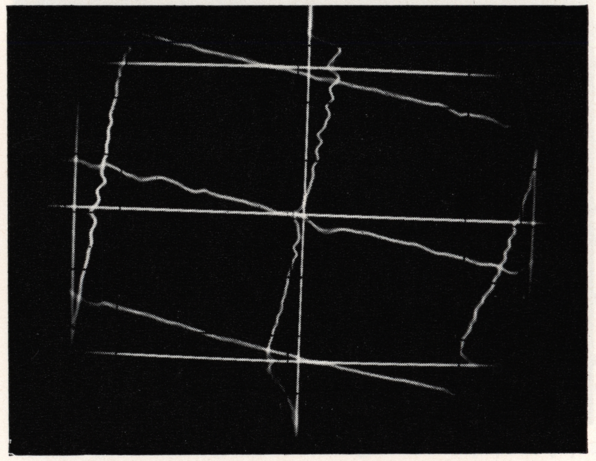

Fig. 20—Effect of tube rotation

Photometric Units: Conversion Tables

We published these tables two or three years ago in another journal. We are now reprinting them because of the growing importance of *light* in many branches of electronics, and also because we believe that a great deal of confusion still exists in the use of photometric units.

There is much to be said for using only the four preferred units given in BS233:1953. (Glossary of Terms used in Illumination and Photometry.) These are the candela, lumen, lux, and nit. For some purposes, however, it may be sometimes convenient to use the stilb and possibly the phot as well.

These units are all taken from the metric system. The foot-candle, foot-lambert, and so on are deprecated. The tables allow easy conversion of all the widely-used non-metric units to metric.

The candela is one of the six units that form the basis of the International System of Units (Conférence Générale des Poids et Mesures, 1960). The other five are the metre, kilogramme, second, ampere, and degree Kelvin.

We hope to publish, within the next few months, a general background article on photometry and colorimetry (insofar as they concern electronic engineers) and on the related questions of measurement in the infrared and ultraviolet regions of the electromagnetic spectrum. It is probably still necessary to say that photometric units do not apply to infrared and ultraviolet radiation. The units for infrared and ultraviolet will be discussed.

Illumination

		lux	phot	foot-candle
1 lux (1m/m^2)	=	1	10^{-4}	$9\cdot29\times10^{-2}$
1 phot (1m/cm^2)	=	10^4	1	929
1 foot-candle (1m/ft^2)	=	10·76	$10\cdot76\times10^{-4}$	1

Luminance

		nit	stilb	cd/ft^2	apostilb	lambert	foot-lambert
1 nit (cd/m^2)	=	1	10^{-4}	$9\cdot29\times10^{-2}$	π	$\pi\times10^{-4}$	0·292
1 stilb (cd/cm^2)	=	10^4	1	929	$\pi\times10^{-4}$	π	2920
1cd/ft^2	=	10·76	$1\cdot076\times10^{-3}$	1	33·8	$3\cdot38\times10^{-3}$	π
1 apostilb (1m/m^2)	=	$1/\pi$	$1/(\pi\times10^{-4})$	$2\cdot96\times10^{-2}$	1	10^{-4}	$9\cdot29\times10^{-2}$
1 lambert (1m/cm^2)	=	$1/(\pi\times10^{-4})$	$1/\pi$	296	10^4	1	929
1 foot-lambert or 'equivalent foot-candle' (1m/ft^2)	=	3·43	$3\cdot43\times10^{-4}$	$1/\pi$	10·76	$1\cdot076\times10^{-3}$	1

Luminous Intensity

Unit: candela Abbreviation: cd

The 'candle power' of a light source (which is assumed to be a point source; though this, of course, is not realisable in practice). For the full definition of the candela, see Refs. 1, 2, and 4.

Luminous Flux

Unit: lumen Abbreviation: lm

The flow of light from the source, related to the solid angle over which measurement is made. One lumen is the luminous flux produced by a point source of one candela within a solid angle of one steradian.

Illumination

Unit: lux Abbreviation: lx

The luminous flux falling on a surface, expressed as luminous flux per unit area. One lux is the illumination produced by one lumen falling on an area of one square metre. Thus one square metre at a distance of one metre from a point source of one candela, has an illumination of one lux.

Luminance

Unit: nit Abbreviation: nt

The luminous intensity of an illuminated or self-luminous surface, usually expressed as luminous intensity per unit area. A luminance of one nit is a luminance of one candela per square metre.

REFERENCES

1. J. W. T. WALSH. **Photometry** (3rd edition). Constable, 1958.
2. G. W. C. KAYE and T. H. Laby. **Tables of Physical and Chemical Constants** (12th edition). Longmans, Green, 1959.
3. British Standard 233:1953. **Glossary of Terms used in Illumination and Photometry.** British Standards Institution, 1953.
4. J. S. PRESTON. **Photometric Standards and the Unit of Light** (Notes on Applied Science, No. 24). HMSO, 1961.

Drive Circuits for Stepping Motors

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Mullard Central Applications Laboratory

Unidirectional and bidirectional transistorised drive circuits for stepping motors are described. Pulse shaper and power supply circuits are included. Stepping speeds of up to 360 steps per second can be achieved.

INTRODUCTION

Stepping motors are, as their name suggests, motors that are designed to rotate by a series of discrete steps, rather than by continuous movement. Among their applications are those listed in Table 1 and illustrated by Figs. 1 to 7.

TABLE 1
Stepping Motor Applications

Counting
- pulse counters (Figs. 1 and 2)
- slow-speed printers
- medium-speed printers

Positioning
- servomotors (digital positioning) for
  - potentiometers (Fig. 3)
  - variable capacitors (Fig. 4)
  - variable transformers
  - valves for fluid control
- stepping selector switch
- digital-to-analogue converters (Fig. 5)
- x-y recorders

Variable programming
- lathes (Fig. 6)
- industrial washing machines
- industrial cookers
- laboratory equipment
- furnaces
- photographic developing baths
- X-ray equipment

Variable drive
- punched tape
- magnetic tape
- continuous chart (Fig. 7)

PRINCIPLE OF MOTOR

The stepping motor has been derived from the conventional synchronous motor. The differences lie in the construction of the stator coils and in the nature of the electrical supply that must be used.

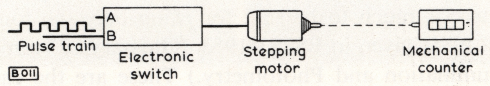

Pulse train
A
B
Electronic switch
Stepping motor
Mechanical counter
B011

Fig. 1—Pulse counter

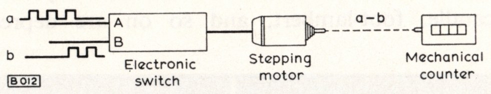

a
b
A
B
a−b
Electronic switch
Stepping motor
Mechanical counter
B012

Fig. 2—Algebraic pulse counter

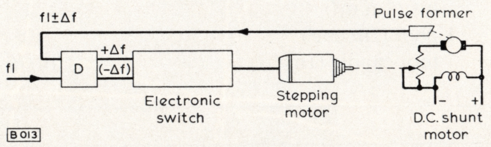

f1±Δf
Pulse former
f1
D
+Δf
(−Δf)
Electronic switch
Stepping motor
−
+
D.C. shunt motor
B013

Fig. 3—Potentiometer control (for speed-stabilisation of a d.c. motor). D is a frequency-differentiator and phase-discriminator

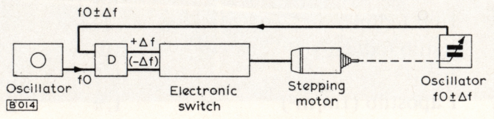

f0±Δf
D
+Δf
(−Δf)
f0
Oscillator
Electronic switch
Stepping motor
Oscillator f0±Δf
B014

Fig. 4—Variable capacitor control (for synchronisation of two oscillators)

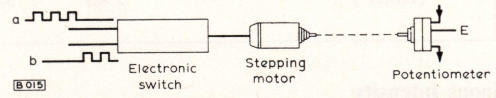

a
b
Electronic switch
Stepping motor
E
Potentiometer
B015

Fig. 5—Digital-to-analogue converter. The voltage E is a function of the number of pulses of (a) and (b)

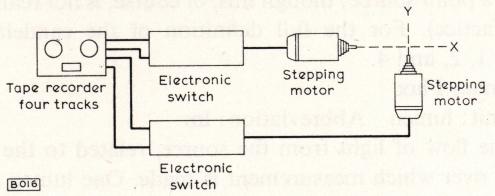

Y
X
Tape recorder four tracks
Electronic switch
Stepping motor
Stepping motor
Electronic switch
B016

Fig. 6—Lathe programming

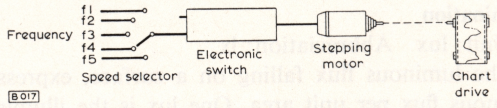

f1
f2
Frequency f3
f4
f5
Speed selector
Electronic switch
Stepping motor
Chart drive
B017

Fig. 7—Chart drive

The motor consists of a permanent magnet rotor with twelve pairs of poles and two stators. Each stator consists of a centre-tapped coil. The stator fields are set up by a direct current flowing through the stator coils. Step rotation is obtained by alternately reversing the field direction of the two stators with a switch. A mechanical switch may be used for low frequency switching; an electronic switch can be used for both high and low frequency switching.

The principles of switching are illustrated by Fig. 8 (in which the switches S_A and S_B may be either mechanical or electronic) and Table 2.

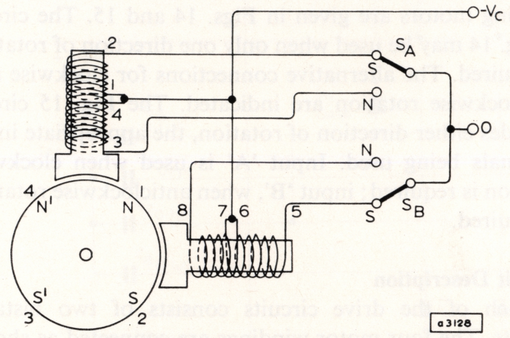

−Vc
2
S_A
S
1
N
4
3
O
N
4
N'
1
N
8
7
6
5
S
S_B
O
S'
S
3
2
a3128

Fig. 8—Switching of stepping motor

TABLE 2
Switching Sequence

Clockwise Rotation Switch position S_A	Clockwise Rotation Switch position S_B	Clockwise Rotation Rotor position N	Anticlockwise Rotation Switch position S_A	Anticlockwise Rotation Switch position S_B	Anticlockwise Rotation Rotor position N
S	S	1	S	S	1
N	S	2	S	N	4
N	N	3	N	N	3
S	N	4	N	S	2
S	S	1	S	S	1

When a mechanical switch is used, the d.c. supply is switched from one half of the coil to the other by means of a single-pole switch to reverse the direction of the field of the stator. Alternatively, switching can be accomplished by means of an electronic switch. The stator coils are then incorporated in two bistable circuits.

Between each switching operation, d.c. passes through the stator coils and supplies a 'holding torque' that holds the rotor in the position to which it had been moved by the preceding switching operation.

When the current is reversed in one of the stators, the rotor is turned through an angle that is determined by the number of poles and the number of stator windings. In motors with twelve pairs of coils for each stator, the number of switching operations for one complete revolution of the rotor is

$$12 \times 2 \times 2 = 48,$$

so that the motor steps 360/48 = 7·5° for each field reversal. The direction of rotation is determined by the sequence in which the stator coils are switched, as shown in Table 2.

MOTOR CHARACTERISTICS

Three types of stepping motor are at present available from Mullard Ltd. Their main features are as follows:

	AU5035/80	AU5055/80	AU5105/80	
Max. working torque	35	100	300	g.cm
Max. no. of steps per second	360	300	150	steps/s
Max. diameter*	35	50·5	44×50·5	mm
Max. height*	21	23·5	75	mm

*excluding fixing lugs, connectors, and spindle

The general appearance of the motors is shown in Figs. 9 and 10, which show the two larger motors just over half-size.

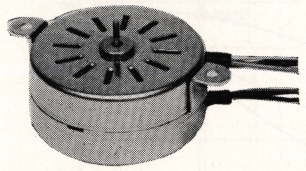


Fig. 9—Stepping motor type AU5055/80

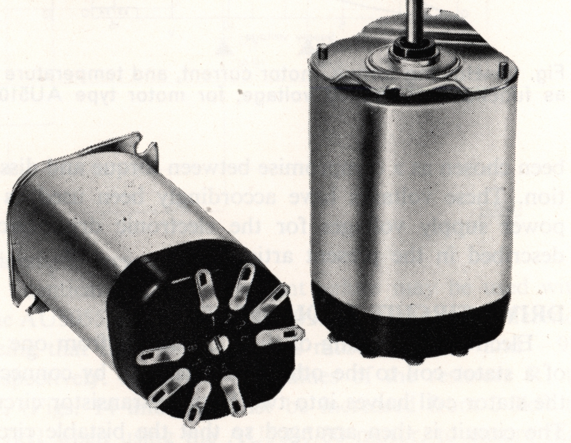


Fig. 10—Stepping motor type AU5105/80

Reduction gears are available for reduction of the angle of rotation per step.

The variation of holding torque against applied voltage is given in Figs. 11, 12, and 13, for the three motors. It will be seen that the increase in holding torque begins to fall at higher supply voltages, whereas the internal temperature rise increases more steeply. The supply voltages, as indicated by the dotted lines, have therefore

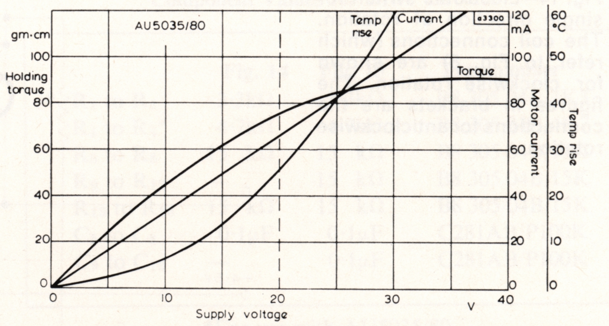


Fig. 11—Holding torque, motor current, and temperature rise as functions of applied voltage, for motor type AU5035/80

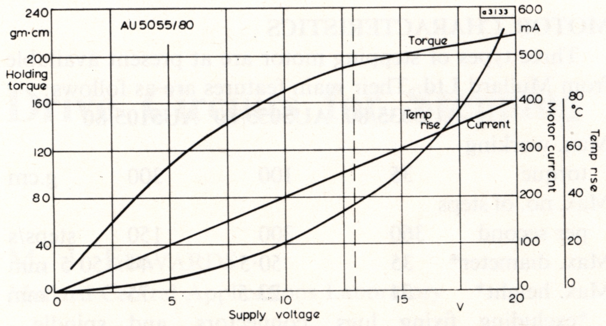


Fig. 12—Holding torque, motor current, and temperature rise as functions of applied voltage, for motor type AU5055/80

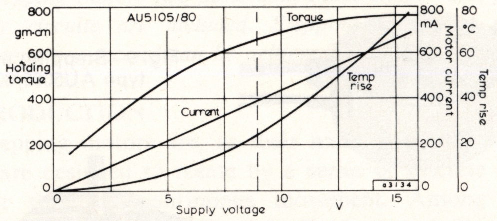


Fig. 13—Holding torque, motor current, and temperature rise as functions of applied voltage, for motor type AU5105/80

been chosen as a compromise between torque and dissipation. These voltages have accordingly been used as the power supply voltages for the electronic drive circuits described in the present article.

DRIVE CIRCUIT REQUIREMENTS

Electronic switching of the d.c. current from one half of a stator coil to the other is carried out by connecting the stator coil halves into two bistable transistor circuits. The circuit is then arranged so that the bistable circuits change state alternately. The two bistable circuits return to their initial state after having been switched four times.

An electronic input pulse is required for each switching, one pulse producing one step of the motor. The sequence of switching the currents in the stator coils defines the direction of rotation. A reversal of the switching sequence will produce rotation in the opposite direction, as already shown in Table 2.

A train of electronic pulses to carry out the switching must be available, and it is typically obtained from the switching of a relay or photocell, or from an oscillator.

ELECTRONIC DRIVE CIRCUITS

Circuits which may be used to drive the three available stepping motors are given in Figs. 14 and 15. The circuit of Fig. 14 may be used when only one direction of rotation is required. The alternative connections for clockwise and anticlockwise rotation are indicated. The Fig. 15 circuit provides either direction of rotation, the appropriate input terminals being used. Input 'A' is used when clockwise rotation is required; input 'B', when anticlockwise rotation is required.

Circuit Description

Each of the drive circuits consists of two bistable circuits. The four motor windings are connected as shown to the collectors of the four transistors TR_1 to TR_4. Since the windings have a high inductance associated with them, the action of switching off the current in any winding will cause a high voltage to be induced in the winding. Diodes D_1 to D_4 short-circuit the induced voltage and prevent damage to the transistors.

The bistable circuits are switched alternately by the four logic gates consisting of R_5 to R_8 and D_5 to D_8 (Fig. 14) or R_5 to R_{12} and D_5 to D_{12} (Fig. 15). The capacitors C_5 to C_8 (Fig. 14) or C_5 to C_{12} (Fig. 15) enable the input pulses to be fed to each logic gate while preventing any direct connection between the four gates.

The value of the input capacitors was chosen so that the input charging time-constant (for example C_5 and R_{13} in parallel with R_7, R_3, and R_{16}) was such that pulses of 5ms rise-time or faster could be accepted. A relay, or a silicon or germanium photocell, may therefore be used to operate the drives directly, as the switching times are shorter than 5ms. A cadmium sulphide cell will be too slow unless light of high intensity is used.

The bidirectional feature of the Fig. 15 circuit is achieved by adding a further set of four logic gates to the circuit of Fig. 14. This addition reverses the sequence of switching the two bistable circuits.

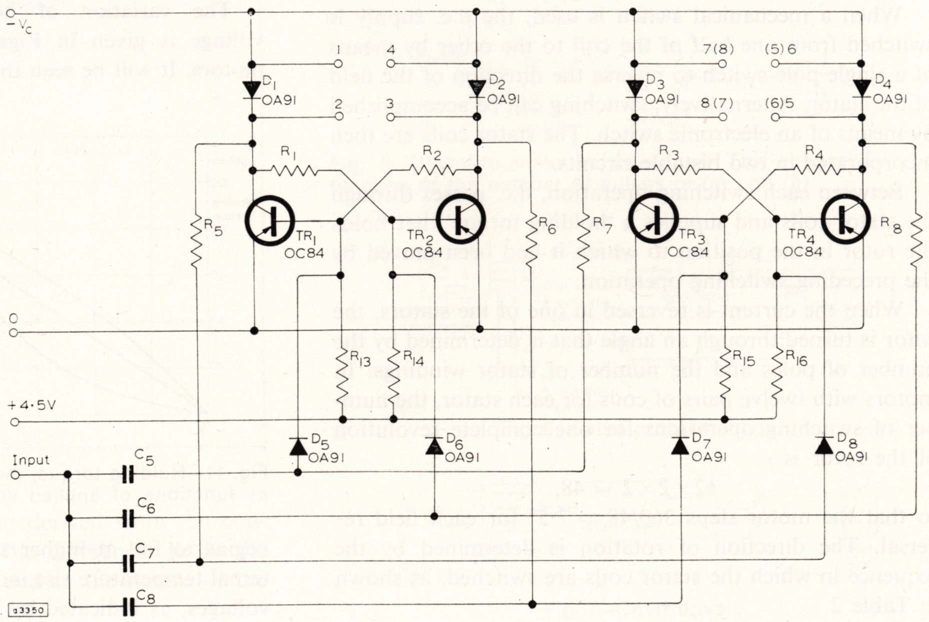


Fig. 14—Electronic switch for single direction of rotation. The coil connections (which refer to Fig. 8) are shown for clockwise rotation. The figures in brackets are the connections for anticlockwise rotation

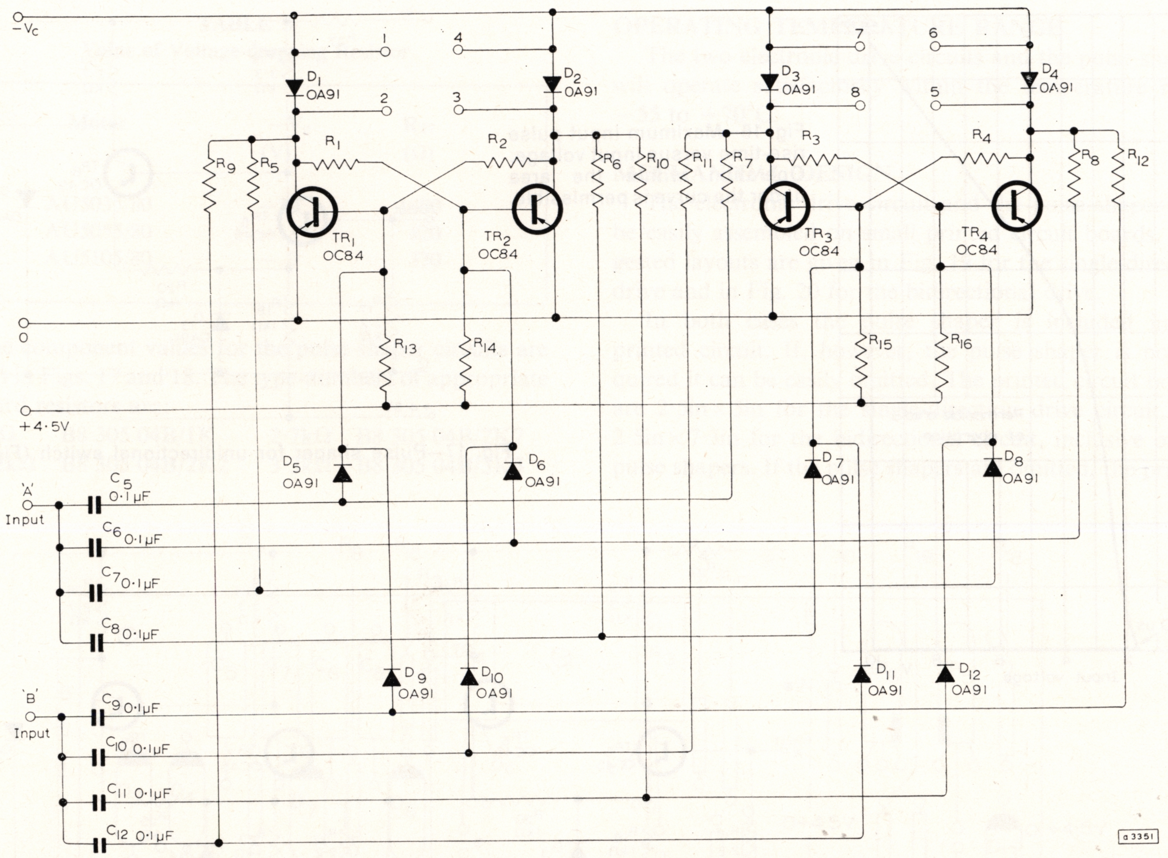


Fig. 15—Electronic switch suitable for both directions of rotation. Input 'A' is for clockwise rotation; input 'B' for anticlockwise

Power Supply Requirements

The operating voltages were determined from Figs. 11, 12, and 13. These voltages were accordingly used as the $-V_c$ power supply for the electronic drive. In addition, a positive 4·5V supply is required. The voltages required, and the currents consumed, are given in Table 3.

TABLE 3
Power Requirements

Motor	$-V_c$ (V)	Current: $-V_c$ supply (mA)	Current: +4·5V supply (mA)
AU5035/80	−20	66	11
AU5055/80	−13	260	9
AU5105/80	−9	400	9

Circuit Values

Identical circuit component values may be used with the AU5055/80 and AU5105/80 motors, the only difference being that the supply voltages must be −13V and −9V respectively. However, the values of the resistors R_1 to R_4 (Figs. 14 and 15) must be increased from 3·3kΩ to 4·7kΩ when using the AU5035/80 motor, to ensure that the transistors can be fully cut off under the most adverse conditions; that is, when operating the electronic drive at high ambient temperatures.

The component values for the two drive circuits are given in Table 4.

TABLE 4
Component Values for Drive Circuits

	Fig. 14	Fig. 15	Type-number
R_1 to R_4	3·3kΩ	3·3kΩ	B8 305 04B/3K3
R_1 to R_4*	4·7kΩ	4·7kΩ	B8 305 04B/4K7
R_5 to R_8	15 kΩ	15 kΩ	B8 305 04B/15K
R_9 to R_{12}	—	15 kΩ	B8 305 04B/15K
R_{13} to R_{16}	15 kΩ	15 kΩ	B8 305 04B/15K
C_5 to C_8	0·1μF	0·1μF	C281AB/P100K
C_9 to C_{12}	—	0·1μF	C281AB/P100K

*For use with AU5035/80

Input Pulse Requirements

The input pulse requirements for the electronic drive circuits are identical, whichever motor is used.

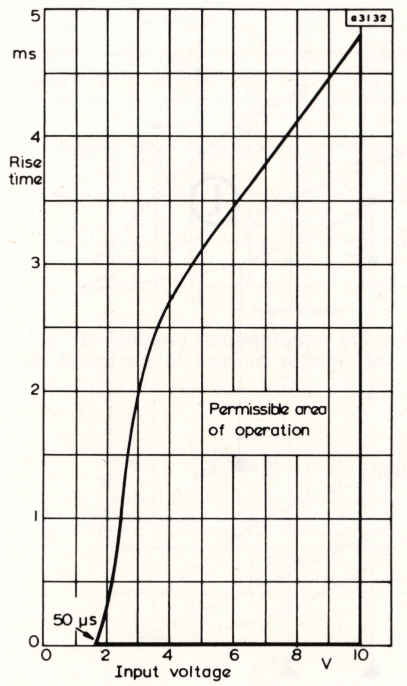


Fig. 16—Maximum input pulse rise-time versus input voltage. Operation within the area under the curve is permissible.

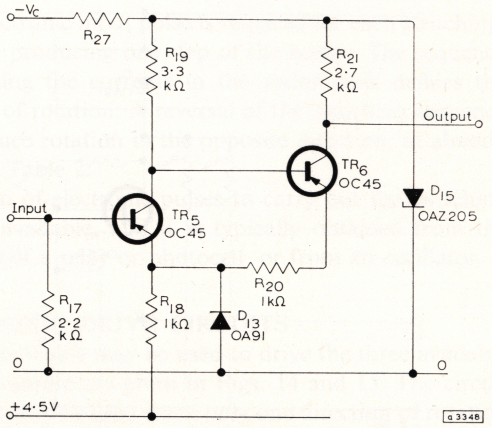

−Vc
R27
R19 3·3 kΩ
R21 2·7 kΩ
Output
TR6 OC45
D15 OAZ205
Input
TR5 OC45
R20 1kΩ
R17 2·2 kΩ
R18 1kΩ
D13 OA91
0
+4·5V
a3348

Fig. 17—Pulse shaper for unidirectional switch (Fig. 14)

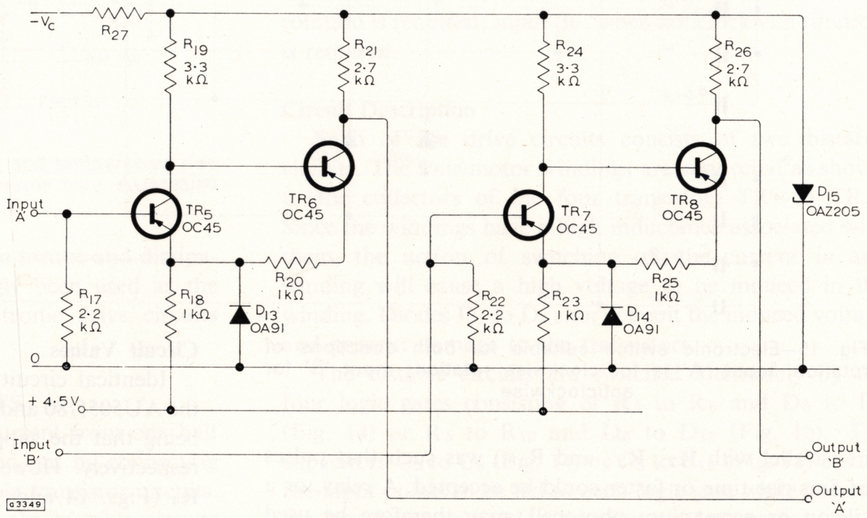

−Vc
R27
R19 3·3 kΩ
R21 2·7 kΩ
R24 3·3 kΩ
R26 2·7 kΩ
Input 'A'
TR5 OC45
TR6 OC45
TR7 OC45
TR8 OC45
D15 OAZ205
R17 2·2 kΩ
R18 1kΩ
R20 1kΩ
D13 OA91
R22 2·2 kΩ
R23 1kΩ
R25 1kΩ
D14 OA91
0
+4·5V
Input 'B'
Output 'B'
Output 'A'
a3349

Fig. 18—Pulse shaper for bidirectional switch (Fig. 15)

Two properties of the incoming pulse are important in determining the operation of the electronic drive circuit; namely, the peak voltage of the input pulse, and its rise-time. The rise-time is taken as the time required for the voltage to increase from 10% to 90% of its peak value.

Fig. 16 shows the longest allowable rise-time for different values of input voltage. It will be seen that a voltage pulse of at least 1·6V must be used. The maximum permitted rise-time for a pulse of this voltage is 50μs. The maximum permissible voltage is 10V, which is the maximum permissible base-emitter voltage of the OC84 transistor. At 10V, the maximum permissible rise-time of the pulse is 4·9ms. Any pulse whose rise-time/voltage intersection lies within the area indicated in Fig. 16 will operate the electronic switch; but the rate of rise must not exceed 1V/μs, otherwise the pulse rise-time rating of the input capacitors will be exceeded.

PULSE SHAPER

If the available pulse does not meet the requirements for rise-time, a pulse shaper must be used. One suggested circuit for a pulse shaper is a Schmitt trigger.

The circuits shown in Figs. 17 and 18, based on the Schmitt trigger circuit, are for a single pulse shaper used with the single-direction electronic drive (Fig. 14) and a double pulse shaper for use with the bidirectional electronic drive (Fig. 15). In Figs. 17 and 18, transistor TR$_5$ is normally conducting and TR$_6$ cut off. A positive voltage at the input will cut off TR$_5$ and switch on TR$_6$. The collector of TR$_6$ goes rapidly positive by 5V and remains at this level until the voltage at the input falls below the triggering voltage level. In this way a 5V positive pulse having a rise-time better than 80μs is obtained. This lies well within the limiting conditions given by Fig. 16.

The pulse shaper input pulse can be of any shape, provided that the voltage exceeds 1V. The input pulses can, if desired, be a sinewave, the positive half-cycles only being used. The output pulse has the same voltage and rise-time whatever the input pulse. To ensure clean switching, the input pulse should be at least 1ms wide at the 1V level. The maximum input voltage is limited to the inverse breakdown base-emitter voltage for the OC45, namely 10V.

The diodes D$_{13}$ and D$_{14}$ help to stabilise the trigger input level at 1V. The zener diode D$_{15}$ stabilises the supply voltage used for the pulse shaper to 7·6V. The additional voltage, V$_c$—7·6V, is dropped across R$_{27}$. Appropriate values for this resistor are given in Table 5.

TABLE 5
Value of Voltage-dropping Resistor

Motor	−V$_c$ (V)	R$_{27}$ (Ω)
AU5035/80	−20	1800
AU5055/80	−13	820
AU5105/80	− 9	330

The component values for the pulse-shaper circuits are shown in Figs. 17 and 18. The type-numbers of appropriate Mullard resistors are:

1kΩ	B8 305 04B/1K	2·7kΩ	B8 305 04B/2K7
2·2kΩ	B8 305 04B/2K2	3·3kΩ	B8 305 04B/3K3

OPERATING TEMPERATURE RANGE

The two electronic drive circuits and the pulse shapers will operate satisfactorily within the temperature range −55 to +70°C.

MECHANICAL DESIGN

The electronic drive circuit and the pulse shaper may be easily assembled on small printed circuit boards. Suggested layouts are given in Fig. 19 for the single-direction drive and in Fig. 20 for the bidirectional drive.

In both cases the pulse shaper is included in the printed circuit. If, however, the pulse shaper is not required it can be easily omitted. The printed circuit boards are 2·5in × 5in for the single-direction drive circuit, and 2·5in × 7·3in for the bidirectional circuit, inclusive of the pulse shapers. If the pulse shapers are omitted, the printed

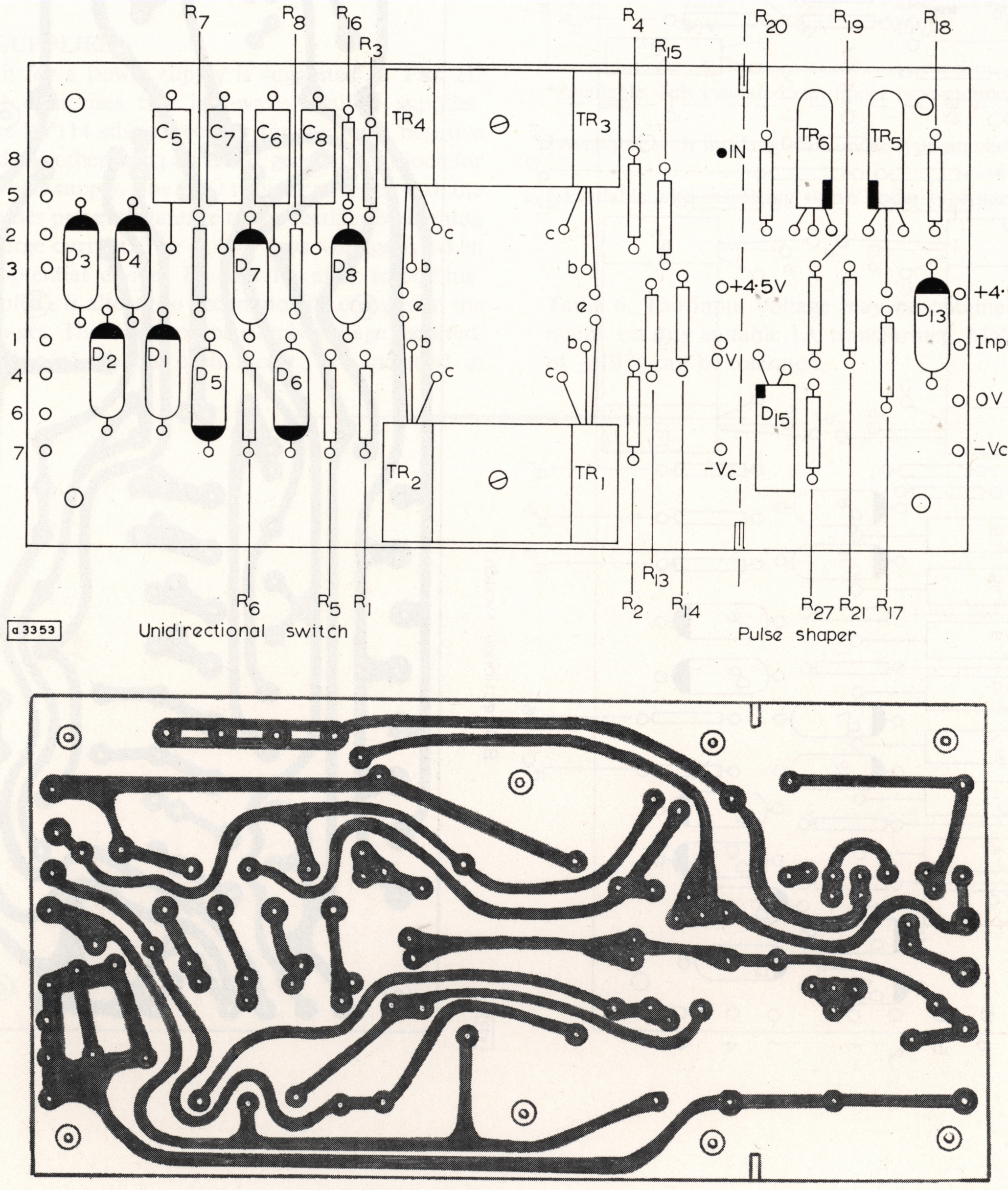

R7
R8
R16
R3
R4
R15
R20
R19
R18
C5
C7
C6
C8
TR4
TR3
IN
TR6
TR5
8
5
2
3
1
4
6
7
D3
D4
D7
D8
D2
D1
D5
D6
c
b
e
+4·5V
0V
−Vc
D13
D15
Input
TR2
TR1
R13
R6
R5
R1
R2
R14
R27
R21
R17
a3353
Unidirectional switch
Pulse shaper

Fig. 19—Printed circuit for unidirectional switch and pulse shaper. Shown actual size; copper side down

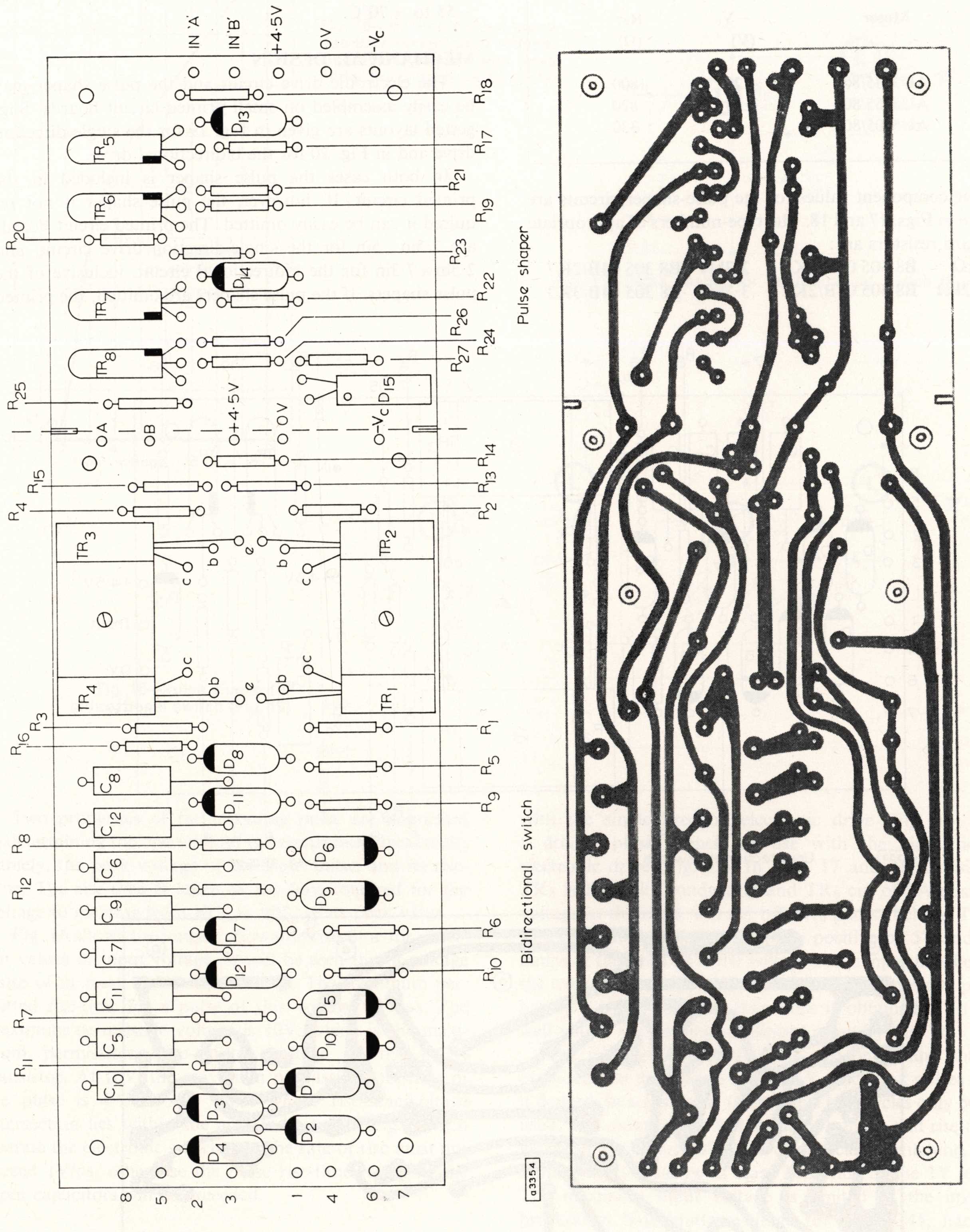
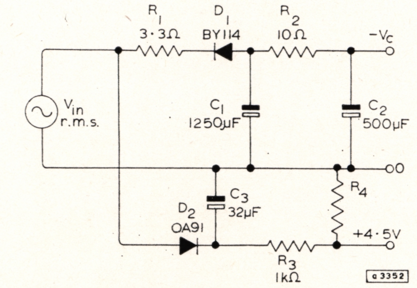

IN 'A'
IN 'B'
+4·5V
0V
-Vc
0A
0B
0+4·5V
0V
0-Vc
R18
R17
R21
R19
R20
R23
R22
R26
R27
R24
R25
R14
R13
R15
R4
R2
R3
R16
R1
R5
R9
R8
R12
R6
R10
R7
R11
TR5
TR6
TR7
TR8
TR3
TR2
TR4
TR1
D13
D14
D15
D8
D11
D6
D9
D7
D12
D5
D10
D1
D2
D3
D4
C8
C12
C6
C9
C7
C11
C5
C10
b
c
e
8
5
2
3
1
4
6
7
Pulse shaper
Bidirectional switch
a3354

Fig. 20—Printed circuit for bidirectional switch and pulse shaper. Shown actual size; copper side down

Fig. 21—
Power supply


R1
3·3Ω
D1
BY114
R2
10Ω
-Vc
Vin
r.m.s.
C1
1250μF
C2
500μF
0
R4
C3
32μF
D2
OA91
+4·5V
R3
1kΩ
a3352

circuit boards are reduced in length by 1·2in and 2·3in respectively. Both the printed circuits when assembled have an overall height of ½in. The 6BA clearance holes are provided for use when it is necessary to secure the printed circuit boards to a base. If the base is of metal, clearance will need to be provided between the base and the printed circuit of about ⅛in.

POWER SUPPLIES

A circuit for a power supply is suggested in Fig. 21. The circuit comprises two half-wave rectified supplies, one using a BY114 silicon rectifier for the main negative supply, and the other using an OA91 germanium diode for the +4·5V bias supply. The 3·3Ω resistor in series with the BY114 rectifier prevents damage to the rectifier by limiting the peak surge current. The +4·5V bias voltage is taken from the potential divider R_3 and R_4 after smoothing. The value of R_4 is altered to accommodate changes in the supply voltages. The values of the supply voltage required, and the value of R_4 for each motor, are included in Table 6. The input voltage may be obtained from 50c/s mains via any suitable 1A transformer. Mains variations of ±10% can be tolerated.

TABLE 6
Circuit Values for Power Supply

	AU5035/80	AU5055/80	AU5105/80	Mullard Type
$-V_c$	−20V	−13V	−9V	
$V_{in(rms)}$	15V	18V	16V	
R_1	3·3Ω 2W	3·3Ω 2W	3·3Ω 2W	B8 305/08B/3E3
R_2	10Ω 2W	10Ω 2W	10Ω 2W	B8 305/08B/10E
R_3	1kΩ ⅛W	1kΩ ⅛W	1kΩ ⅛W	B8 305/04B/1K
R_4	390Ω ⅛W	470Ω ⅛W	470Ω ⅛W	B8 305/04B/390 or B8 305/04B/470
C_1	1250μF 40V	1250μF 40V	1250μF 40V	C431/BB/G1250*
C_2	500μF 40V	500μF 40V	500μF 40V	C431/BB/G500†
C_3	32μF 40V	32μF 40V	32μF 40V	C426/AE/G32§

*Available with insulated can under type-number C431/BR/G1250

†Available with insulated can under type-number C431/BR/G500

§Available with insulating sleeve under type-number C426/AM/G32
